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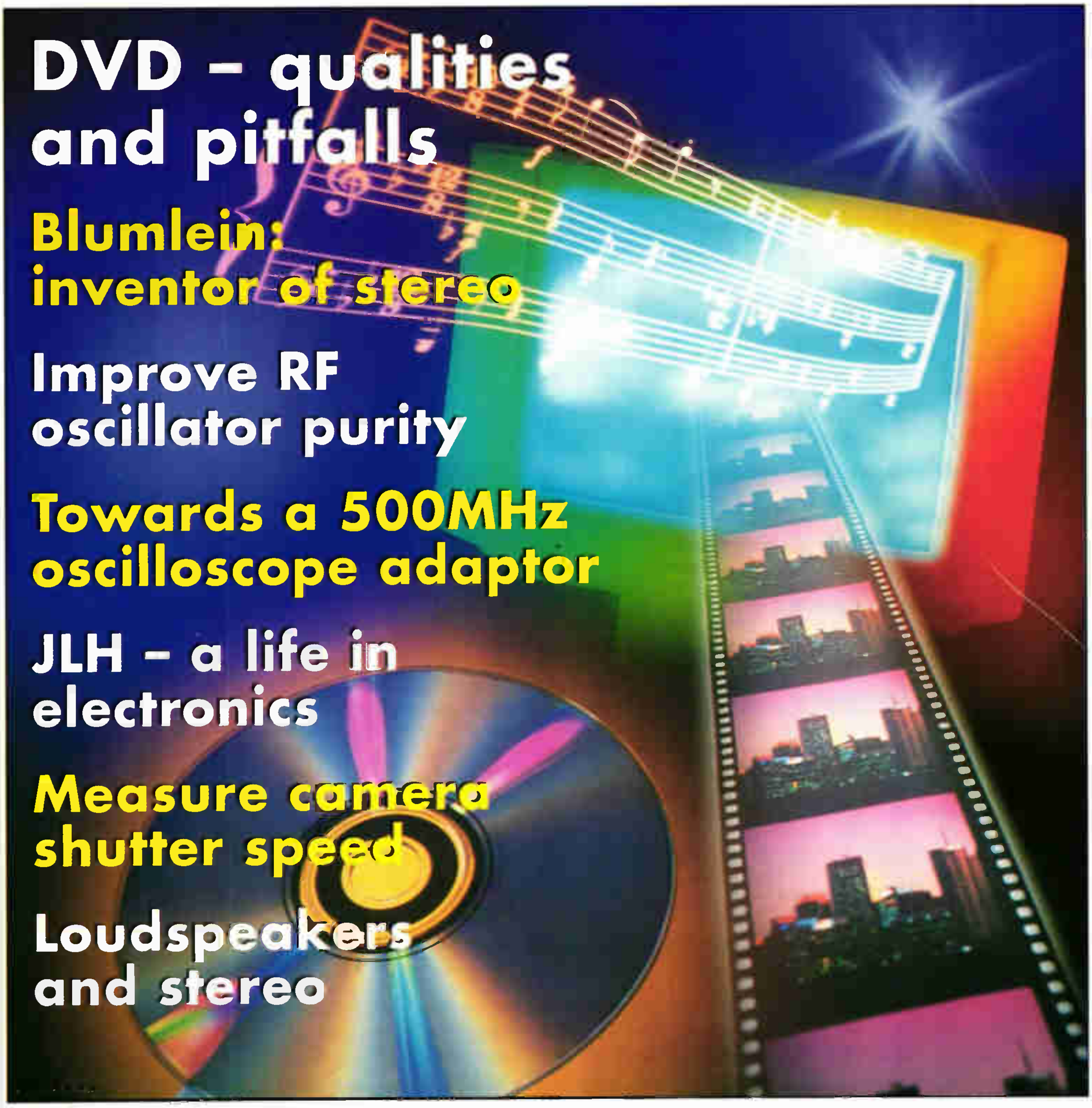

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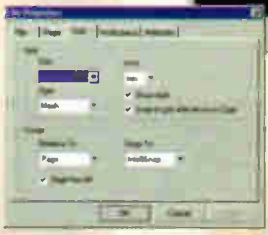

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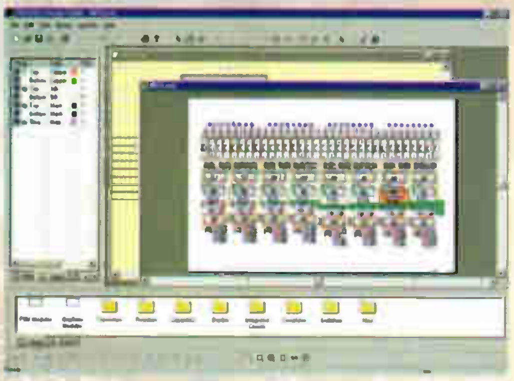

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Society's scientists. Post Y2K.

202 TOWARDS A 500MHZ SCOPE ADAPTOR
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Current conveyors offer a simpler and more elegant solution than conventional op-amps in many analogue circuits. John Lidgey and Khaled Hayatleh illustrate how to reap their full benefits.

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252 ALAN DOWER BLUMLEIN
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256 WEB DIRECTIONS
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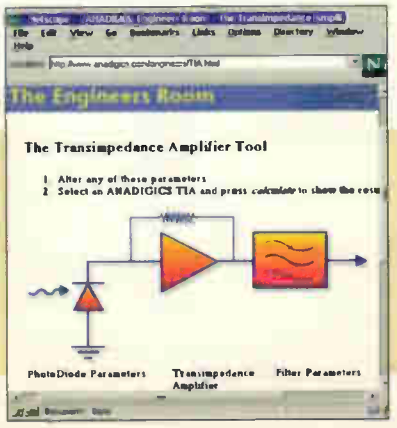

The Engineers Room

The Transimpedance Amplifier Tool

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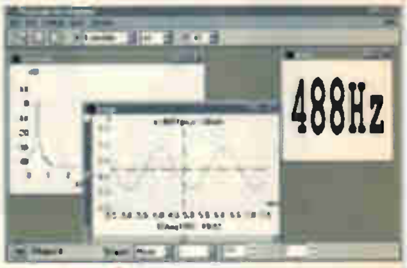
2 Select an ANADIGICS TIA and press *calculate* to show the resu



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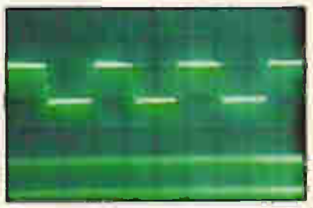

Sampling adaptors have been used in the past to enhance the performance of slow oscilloscopes. According to Ian Hickman's investigations, starting on page 202, modern components promise a 500MHz sampling adaptor.

Following Blumlein's death, his work was shrouded in secrecy. He received neither obituary nor tributes. Read about Blumlein's life and works on page 252.

April issue on sale 2 March

While investigating the benefits of transimpedance over transconductance when designing photodiode amplifiers, Cyril Bateman found this web site that allows you to experiment with components on line. Find out more on page 210.

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EDITOR
Martin Eccles
020 8652 3614

CONSULTANTS
Ian Hickman
Philip Darrington
Frank Ogden

EDITORIAL ADMINISTRATION
Jackie Lowe
020 8652 3614

EDITORIAL E-MAILS
jackie.lowe@rbi.co.uk

DISPLAY SALES EXECUTIVE
Linda Payne
020 8652 3621

ADVERTISEMENT E-MAILS
linda.payne@rbi.co.uk

ADVERTISING PRODUCTION
020 8652 3620

PUBLISHER
Mick Elliott

EDITORIAL FAX
020 8652 8111

CLASSIFIED FAX
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No defence for privatisation

Margaret Thatcher set the government's privatisation ball rolling in the 1980s with successful sell-offs at BT, British Gas and British Airways. John Major in the 1990s turned to the railways and bus services with questionable success.

Now Tony Blair's New Labour is determined to keep up the sell-off tradition. But with their options limited – all the former government-owned services and industries with the highest commercial value have already been sold – the only privatisation target they could find was the Ministry of Defence's advanced, and in many respects top secret, technology research organisation.

It was a hair-brained scheme that would have put strategically important and highly secret research facilities like Malvern and Aldermaston under partial or even total private ownership.

Thankfully the whole sorry affair has led to an embarrassing U-turn by the government.

Nevertheless, the Treasury seems determined to wring privatisation cash out of the Defence Evaluation and Research Agency, known as DERA, come what may. Instead of total privatisation of DERA, the Whitehall mandarins have come up with a plan to split the Agency down the middle; keeping the secret bits under government ownership and selling off the rest in a share sale.

Not surprisingly, the DERA workforce is fearful of job cuts and who can blame them, given the all too familiar necessities of the commercial world.

But what sort of cash does the Treasury expect to raise through a partial privatisation of DERA and its scientists?

Surely the payback to the government will be minuscule compared to previous share sales at BT and BA?

Consider that the auctioning of the next round of mobile phone licences this summer will net £2bn plus for the Treasury. So why bother with DERA particularly as the privatisation plan is being criticised by Ministry of Defence (MoD) staff and is already putting strain on Anglo-US relations.

The MoD's chief engineers claim the proposed sell-off and break-up of various defence research facilities now grouped together as DERA could pose a threat to the country's strategic defensive capability.

Strong words indeed, but it seems they may have a point. The UK's closest military partner, the US, has been unhappy about the proposals from the outset. It is widely thought that it was the objections of the US military that finally forced the government to scrap its original wholesale privatisation plan last month.

It seems likely, however, that the Pentagon, just like the DERA scientists, will be less than satisfied with the fudge which the MoD has now come up with.

After the proposed split and partial privatisation DERA's management will have the freedom to seek out lucrative research contracts wherever they are in the world. Indeed, to expand the business and generate a return for investors they will be obliged to look overseas for more and more work. It is feasible that defence research groups at Malvern will be carrying out more work for foreign governments than they will be for the MoD.

Security could be compromised, and that is what is worrying the US military.

US Under Secretary for Defense Jacques Gansler has already made it clear to the UK government that it could no longer share military technological secrets with a commercialised Agency.

So what does DERA do which is so important to the UK? It takes part in collaborative research with UK defence firms like British Aerospace. A recent example of such a collaboration is the Pathfinder radar software programme, which reduces the cost of naval surveillance by allowing the radar processing software to run on commercial computer systems.

For an example of DERA's strategic military importance simply look a the work of the agency's ordnance researchers in developing for the MoD a fully global positioning system (GPS) guided artillery shell.

Technology like this does not come cheap and investment in DERA is a major drain on the government's defence budget. So help from the private sector would be an advantage, but there is a commercial price to pay and in the case of a strategically important agency like DERA. If that price involves breaking up the agency and privatising chunks of its activities then that price is just too high.

The government has been steadily reducing its financial commitment to defence research since its 'Options for Change' initiative which followed the fall of the Iron Curtain over 10 years ago.

What remains within DERA is an efficient research organisation employing 12000 staff with a budget of £1bn which cannot benefit from further dismemberment.

The government's new fudged privatisation plan will attempt to raise cash for the treasury by selling shares in the most commercially attract part of DERA, allowing it to operate in the private sector. While preserving the Agency's essential strategic character and protecting the MoD's interests.

Surely, it cannot work? ■

Richard Wilson

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Subscriptions: Quadrant Subscription Services, Oakfield House Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 01444 445566. Please notify change of address.
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UPDATE

UK could be world leader in digital TV

The UK has a unique opportunity to lead the world in the production of advanced digital TV technology, according to Europe's leading set-top box manufacturer.

Pace Micro Technology's CEO Malcolm Miller said: "The UK has the opportunity to lead the world in digital TV, just as the US has done with the PC and the Internet."

The future of digital TV in the UK will soon include e-mail, Internet and intelligent programming.

Another UK company, Two Way TV, is looking to become a leading provider of digital interactive TV content. The company recently formed a joint venture with US firm Interactive Network to deliver real-time interactive games and sports channels.

Two Way TV's managing director Simon Cornwell, said: "Through the venture, we will be able to form partnerships to produce live interactive TV entertainment for distribution on digital satellite, cable and broadband Internet platforms.

With key patent protection, the venture will be in a unique and powerful position in the rapidly developing interactive TV market."

Pace has grown to become the number one set-top box maker in the UK and number 4 in the world. The firm took on 120 engineers this year and plans to take on 150 more next year, 80 per cent of which will be in the UK. The firm is valued at around £1bn today, growing from £100m two years ago.

"We are now showing a box with an embedded hard drive and 20Gbyte storage space. It will be able to download programmes for you, working out what kinds of things you watch," said Miller.

The technology has the potential to replace "the boring part of life," with the provision of home banking services and home shopping, added Miller.

System suppliers are benefiting from the highly competitive UK digital TV market, where satellite, terrestrial and cable broadcasters are all vying for market share.

Alex Mayhew-Smith *Electronics Weekly*

Dyson's delighted that robots suck

High flying domestic appliance firm Dyson is going hell for leather in its use of electronics.

"Electronics is absolutely vital," says James Dyson, chairman of Dyson Domestic Appliances. "It may be increasing cost but it is improving performance and enhancing the user interface."

Dyson was talking at the launch of the firm's first autonomous vacuum cleaner. The £5m development involving 40 designers is its first product with a large electronics content.

The company also hinted at other products under development – not vacuum appliances – which also use significant electronics.

Dyson expects to recruit over 400 people in the coming year, and of these up to 100 will be electronics engineers.

"We want to attract electronics people in a big way," he added.

The robotic vacuum cleaner will be on sale from spring costing £2500.

"The move to robotics – especially vacuum cleaners – will happen very quickly, in five years or less," said Dyson. "We have got the technology so we can supply it now. People do not want to vacuum if they can avoid it."

However, the main reason for the company's uptake is an increasing shortage in service personnel, he said.

The company faced two challenges in the robot's design: the battery powered cleaner had to pick up at least as well as a mains powered one does, and its coverage of floor space had to improve on conventional human-operated cleaners.

"Ours is more intelligent than a human: it is methodical and it knows where it's been," he said.

Roy Rubenstein *Electronics Weekly*

ROBOT VACUUM CLEANER AT A GLANCE

- Fast charge nickel-cadmium batteries are used, providing the robot with a 30 minute working life
- It uses four micro-controlled switched-reluctance motors for the front wheels, suction and brush bar
- It has three on-board computers and over 50 sensors including infra-red, ultrasonic and light compass ones
- The unit pauses for 20s if kids or pets interact with it

Wireless Internet trials imminent

A trial of third generation (3G) wireless Internet technology by Vodafone and Nortel Networks is due to start early this year in London. The trial will last until at least March, but the number of participants has not been disclosed.

"We hope to gain a clearer understanding of the technology and consumer behaviour of wireless Internet and use this to guide us in deploying 3G networks worldwide," said Alan Harper, Vodafone managing director.

Nortel and Panasonic will provide the radio equipment, core data network and terminals. A high-speed IP data connection will connect trial equipment to corporate intranets and the Internet.

Prototype terminal equipment will be used to test various data and voice services. These devices will include a mobile phone with built-in camera and video screen, which uses the MPEG4 video codec and can transmit and receive data at up to 64kbit/s.

Other devices will include lightweight, pocket-sized voice terminals, wireless modems for laptop computers, and a mobile data device with transmission speeds up to 384kbit/s.

Promotion for academic links is inadequate, says MP

An MP has called on the government to set up an agency to oversee the promotion of links between high-tech business and universities.

"At the moment there are many small and not so small schemes for universities and businesses," explains Ian Pearson, MP for Dudley South. "All the surveys show that this is confusing. We should have a first-stop shop that would take over these schemes."

Pearson hopes his idea will be included in the government's science and innovation white paper, due in February or March.

It has already been submitted to Stephen Byers, the trade and industry secretary, and according to Pearson was described by Byers as being "controversial and thought-provoking".

Pearson's paper will be published early next year by the Social Market Foundation.

This week the Department of Trade & Industry said it planned to secure a 50 per cent increase in the number of spin-out companies from universities.

"The UK is not making the most of the commercial potential of its world-class academic research," the DTI said in a report UK Competitiveness Indicators 1999, which is the first in a series intended to gauge the success of the UK as a knowledge driven economy.

The report also said there had been a big increase in the amount of venture capital money available to early stage start-up businesses. The amount available grew from about £150m in 1997 to around £300m last year.

Caught on camera... Bond girl Denise Richards is seen in the latest 007 film, making use of Hewlett-Packard's 430se hand-held Windows CE terminal.

BT's broadband lead claim in question

BT boss Sir Peter Bonfield's contention that the UK will have a world lead in broadband access looks increasingly unachievable. South Korea, for one, looks like beating the UK.

Korea's three telecoms operators, Dacom, Hanaro and Korea Telecom – have already started offering broadband access to consumers in the country's major cities for a mere £12 a month – about one third the price BT is expecting to charge when it gets its service up next Spring.

The US is in the lead on broadband access with some 1m lines, according to analysts Dataquest, and Singapore is thought to be in second place. BT says it expects to have 400 exchanges ready to offer the technology in the Spring – but it wants to charge so heavily for the service that it may be hobbled.

Samsung is now offering a UDSL (UltraDSL) chip-set delivering 12Mbit/s for telephone lines and is developing a cable modem chip set to provide broadband access for those with cable.

Trench memory transistors promise 25% die size reduction

Infineon Technologies and IBM have extended their DRAM alliance until 2002 as they work on commercialising their revolutionary new trench transistor technology.

In the past, trenches have been used for burying capacitors in the silicon substrate but not, before now, for burying transistors. The IBM/Infineon technology now allows the entire memory transistor array to be buried beneath the surface of a chip, allowing logic circuitry and five or six layers of interconnect to be built on the planar surface.

According to Infineon, this approach reduces die size by some 25 per cent compared with the alternative stack cell technology of building transistors on top of themselves up from the chip's surface.

It can also offer advantages in embedded DRAM because the surface logic and interconnect is on the same level as other parts of a system-on-chip IC.

"This approach allows very good optimisation of logic," points out Dr Willy Beinvogel, a senior v-p in Infineon's memory division.

Beinvogel said that the target for commercialising the trench transistor technology is to use it in the gigabit DRAM at the 0.11μm geometry level for volume production in 2002.

"We are making an intensive effort to get to 0.11μm," said Beinvogel. His preferred route is via 193nm wavelength UV processing, but Infineon is also involved in the ion-beam processing being pursued by IMS of Vienna and is about to join the extreme-UV consortium which has been mainly driven by Intel.

Toshiba has been involved in the IBM/Infineon alliance at the 256Mbit level but ends its interest in the first quarter of 2000.

IBM is ramping down its internal manufacturing of DRAM, but is continuing to develop DRAM technology with the intention of licensing it.

David Manners *Electronics Weekly*

Over 300m mobile phones Internet ready by 2003

The world will have 500 million mobile phone subscribers by 2003 and three quarters of the phones will be Internet enabled.

This prediction by industry association the Wireless Application Protocol (WAP) Forum is backed up further by research from US analysts Gartner Group.

It forecasts that over 95 per cent of new mobile phones shipped in 2004 will be WAP-enabled. WAP is a *de-facto* standard that provides a means of delivering Internet data to mobile terminals.

The analysts also said that annual shipments of portable Internet-enabled devices would exceed those of PCs in 2003.

The first WAP enabled phone from Nokia was launched in the UK by Orange last month and more will follow. The WAP Forum's membership includes handset manufacturers which have committed to shipping WAP-enabled devices. These companies represent 95 per cent of the world market.

Version 1.2 of the WAP specification has just been released by the WAP Forum and its membership now exceeds 250, with 52 of those members joining since September last year.

Gartner Group analysts also predict mobile phone penetration in Europe will exceed an average of 60 per cent of the population by 2004 and that more than three quarters of new hand sets in 2004 will be Bluetooth-enabled.

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CIRCLE NO.108 ON REPLY CARD

PCB manufacturing clearing house on the Web

An e-commerce Web site to help companies get their printed circuit boards made has been launched by Info Elec.

The Ashford company is the second UK firm to use the Web to take the leg work out of getting quotations for PCB work, joining the company PCBQuotes.

Info Elec's approach is to use PCB manufacturers worldwide rather than just those in the UK.

"We have signed up 50," said Olivier Cadic, the firm's chairman. "The aim is to have 900 PCB manufacturers signed within 18 months."

Worldwide there are approximately 4500 PCB makers.

The company checks each manufacturer's suitability before signing them. Quotes for work are then sent to those companies with the required capabilities.

Only when the quote is accepted is the identity of the manufacturer revealed. All the remaining PCB makers are informed of the terms of the winning quote and the country of the manufacturer.

The service is free to buyers. It is also free to manufacturers until they receive a successful quote. Then they pay a 3 per cent charge to Info Elec, which guarantees their payment 30 days after the boards are delivered.

"We help the purchaser find the best location for the work and help the market to be regulated," said Cadic.

http://www.pcb007.com/

Roy Rubenstein

Tiny Linux for embedded systems

Network terminal designer Amino Communications claims to have the world's smallest Linux systems. The systems, on PCBs measuring 49 by 98mm, are aimed at developers of embedded systems such as seat-back entertainment systems, as well as kiosks, web phones, network-centric terminals and set-top boxes. The Linux operating system is particularly suitable to a small footprint as it can run in 2Mbyte of memory. Power consumption is 2W or less, says the firm.

Linux Wall St trading record

An operating system firm beat the record for the biggest first day of trading on Wall Street last week. VA Linux Systems' shares rose 733 per cent in the first day. Shares for the Linux operating system were $30 at the start of trading on Thursday, rising to $250 by the end of the day. The previous record was held by Theglobe.com, an Internet firm, which grew 606 per cent on its first day.

Smart grant challenged

A Labour-dominated committee has clashed with the government over the Smart high-tech grant scheme. In March 1999, the DTI announced an extra £26m increase for the scheme saying £100m would be available for new Smart projects in the next three years. But the committee said the 1999 report gave no forward figures for funding beyond £20.4 million for 1999-2000.

CMOS TV tuner cuts up to 200 components

Broadcom claims to have introduced the world's first CMOS TV tuner that could have the potential to dramatically reduce component counts and shrink the size of TV tuners.

The Broadcom BCM3400 family of tuners will allow manufacturers to use a single tuner chip to replace current designs that use as many as 200 components. This will lower manufacturing costs and improve the reliability of TV tuners. The company has begun sending out first samples of the chip.

"With this level of integration, manufacturers can employ one or more tuners in a wide range of consumer devices such as televisions, digital video recorders, cable modems and set-top boxes," said Allen Leibovitch, analyst with US market research company IDC.

"Our CMOS television tuner brings new levels of cost benefits and integration to one of the last markets dominated by discrete RF component-based solutions," said Broadcom CEO Henry Nicholas.

Plans to quadruple Rambus throughput

Memory design firm Rambus has outlined its technology roadmap for the coming years, which includes a quadrupling of data rate by next year.

The firm is also designing its memory system into communications products such as high-speed data switches and routers.

"Next year we expect to release technology that will permit a further doubling of per-wire data rate," said Dave Mooring, recently promoted to president of Rambus. This will take the transfer rate to 1.6GHz.

"Secondly," Mooring said, "Rambus will quadruple data rate to 6.4Gbyte/s". Markets that could benefit from this include graphics and multimedia.

The move into communications systems could be lucrative. While it only accounts for nine per cent of chip sales, the average selling price per part is higher and there is always potential for Internet growth.

The firm dismisses the challenge posed by double data rate (DDR) memory. "We believe that DDR offers little performance improvement over existing SDRAM," said Geoff Tate, CEO of Rambus. Gains could be as low as 20 per cent, he said. "We think DDR is behind us in performance."

DDR could also infringe on Rambus' 70 patents, Tate thinks. The firm has got hold of DDR samples and is looking to see if they infringe.

Rambus has also set up a new group with the task of managing acquisitions and investment in other companies.

Boom expected for in-car electronics

Electronic content is predicted to reach more than 30 per cent of the value of high-end luxury cars by 2005, according to the Economist Intelligence Unit.

These cars are likely to be in the £20000 to £30000 price bracket at today's prices, but even the more popular three-door hatchbacks will see 20 per cent of their value made up from electronic systems by 2005, says the report.

Power-train electronics, such as engine management systems, will make up 30 per cent of the car's electronics with safety systems accounting for ten per cent.

By far the biggest factor will be the growth of in-car multimedia systems and mobile communications systems. These will account for over half the electronics in cars within five years, says the report.

The report lists features it expects to see fitted to cars within five years. These include Bluetooth wireless networks, biometric access systems, and neural-network-based ignition systems.

With the growth of electronic systems, car manufacturers are moving quickly to 42V power systems to replace traditional 12V systems.

Leak detector wins

Bolton Institute has won an engineering excellence award for its collaboration with Andel of Huddersfield to design a liquid leak detection system. The award from TCS, formerly known as the Teaching Company Scheme, recognises the work Bolton has done to bring microelectronics to firms that do not have the capability for electronic design. Bolton carries out its work through the DTI's Electronics Design initiative and before this through the Microelectronics in Business campaign. ■

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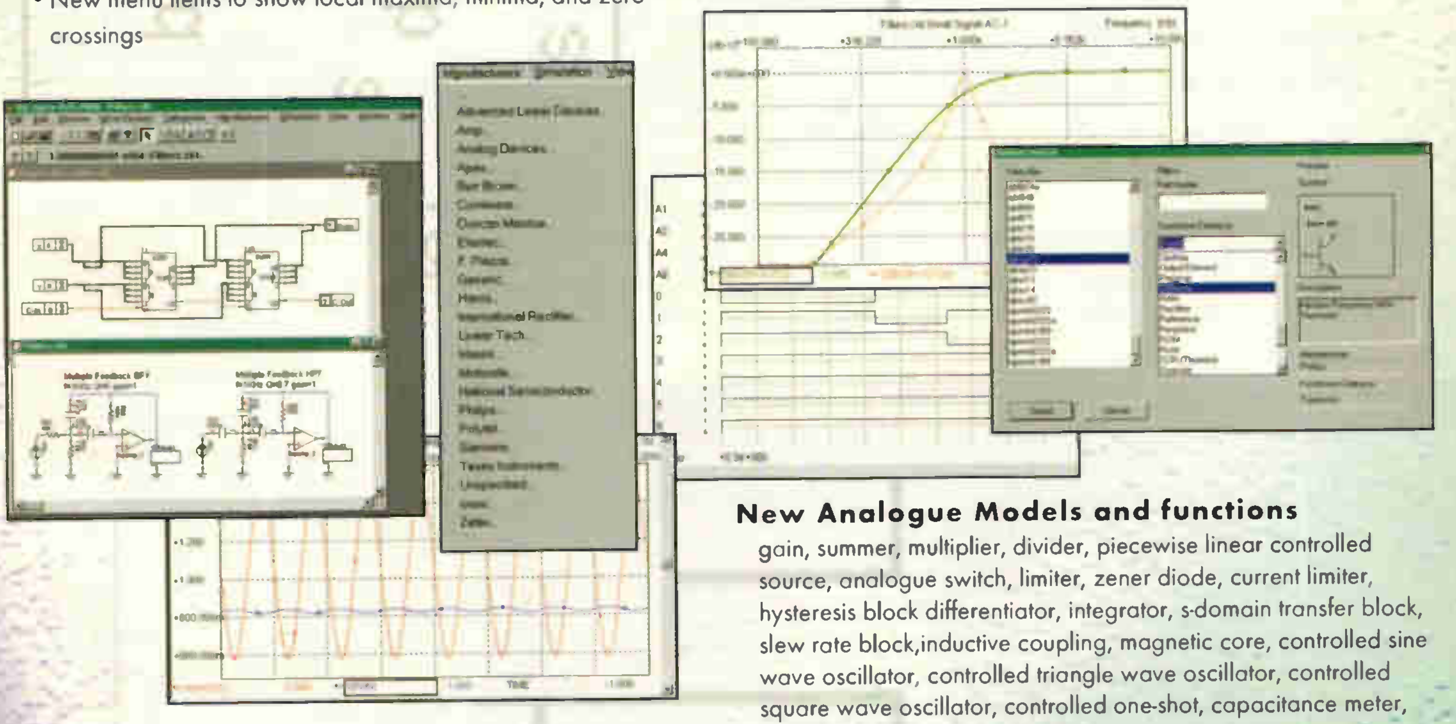

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CIRCLE NO. 109 ON REPLY CARD

Inside DVD

What looks like a CD but holds seven times as much digital data? The answer's a DVD but please don't call it a digital video disc. The many-faceted DVD now has far more capabilities than just video and has shed its old tag in favour of 'digital versatile disc'. Let Andrew Emmerson acquaint you with DVD's qualities – and its pitfalls.

The first significant software product on DVD is Microsoft's Reference Suite 2000 – the latest version of Encarta – from which this volcano was extracted.

It's a tribute to DVD's affordability that this format has taken off so rapidly. With customer acceptance running faster than videotape, laserdisc, and CD, the DVD format has successfully established its own niche as an entertainment and data storage medium.

What's more, it has achieved this at a time when other technologies exist to serve the selfsame functions that DVD fulfils. So what is DVD and what is its attraction?

Superficially, DVD is just a variant of the highly successful compact disc (CD) format first devised for digital audio records by Sony and Philips in 1980. The success of this medium and the realisation that it was capable of handling greater recording densities led to a joint decision between the entertainment, consumer electronics and computer industries to develop a video variant, known then as the digital video disc.

Technical standards were finalised in 1995 and a year later DVD video players were launched in the USA. In 1998 the DVD was reincarnated for data storage as the DVD-ROM.

DVD versus CD

Primarily, DVD differs from audio CDs in packing data more densely to achieve significantly greater data capacity, **Fig. 1**. The pits read by laser beam are smaller, the tracks are more closely spaced and the red laser employed has a shorter wavelength for higher resolution.

The error-correcting system is more robust too. As a result standard DVDs can hold 4.7Gbyte of data – some seven times more than an audio CD, while dual-layer DVDs can increase this to 8.5Gbyte, or 12 times as much, **Fig. 2**.

Video DVDs can hold movies of up to 135 minutes duration on a single layer disc, while using two layers, the Fox/Paramount release of the film *Titanic* manages to cram a film over three hours long onto a single disc.

Their huge capacity makes DVDs ideal also for multimedia applications, offering higher resolution pictures (wide screen or academy format) and more channels of Dolby digital sound. Interactive computer games benefit from richer graphics and high-quality full-motion video, while computer users can store massive databases onto DVD.

Being a disc, the DVD provides virtually instantaneous random access to any block of data and being optical, the disc can be replayed repeatedly without worries about wear and tear.

Another advantage is the higher read speed of the DVD system, with higher data transfer rates than the fastest CD-ROM drive. In addition, the DVD format is an industry standard, supported by the world's leading hardware manufacturers and software providers.

All told, it sounds like perfection at last, with sweetness and light pervading in every direction. Chance would be a fine thing, however, as will be revealed in due course.

Video versatility...

For those accustomed to VHS recordings, the capabilities of DVD offer the same kind of enhancement that CDs have over compact cassettes – summed up as breathtaking picture quality and stunning sound.

Even the most hardened cynic will not fail to appreciate the startling brilliance and clarity of the pictures delivered by DVD, with saturated, noise-free colours and roughly 500 lines of horizontal resolution – more than twice the 240 lines commonly achieved by standard VHS tapes.

Colours are properly saturated, with very little picture noise – even on freeze-frames. On NTSC material the difference between watching a VHS tape and a DVD is quite remarkable and the DVD format has certainly closed the gap between NTSC and PAL. Indeed, it is true to say that the untrained eye cannot often distinguish the two when watching DVD-sourced films.

Digital surround sound with six discrete channels of audio offer a movie theatre-style experience in the viewer's own home, while many DVD players offer a variety of viewing options, including a choice of different languages or subtitles, screen aspect ratios and on some software, up to nine different camera angles. And because most software is divided into chapters, viewers can resume play at the point in the movie where they stopped.

...and data diversity

DVD's increased storage capacity and speed performance relative to CD-ROM has made this medium ideal for software houses and multimedia applications.

A development is DVD-RAM, which is the name given to a rewritable double-sided compact disc providing considerably greater data storage than today's CD-RW systems. There is also, however, a competing standard called DVD+RW. The former standard supports 2.6Gbyte per disc side while DVD+RW handles 3Gbyte per side.

How DVD works

As already indicated, DVD uses a laser to

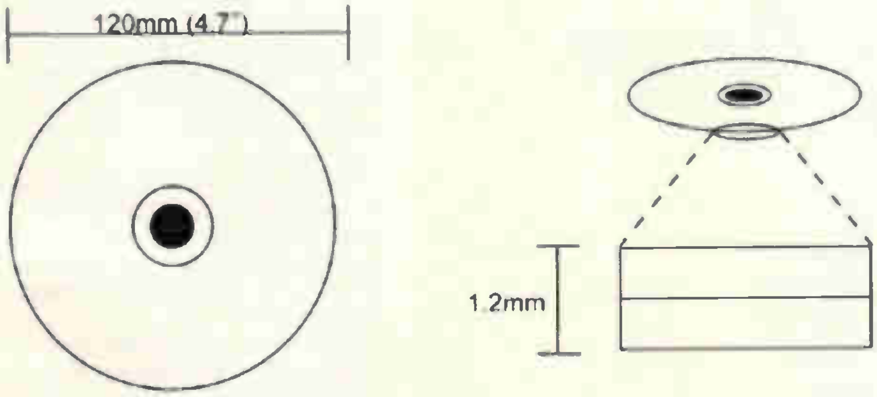


Fig. 1. DVD discs are the same size as CDs, but there is also a two layer option that's twice as thick.

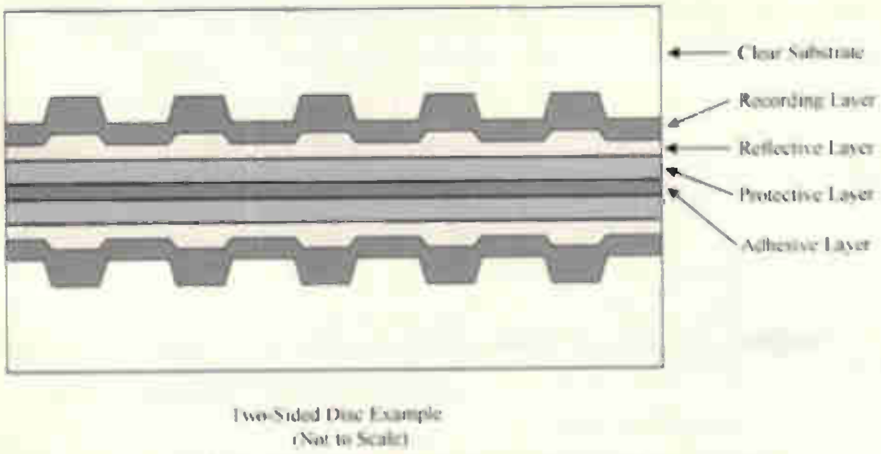


Fig. 2. Layers within layers. Not only can DVD discs be stuck together back-to-back to form a double-sided disc, but they can also have a dual data layer, each layer being read by a different coloured laser. This quadruples the data density of a normal single-sided, single-layer DVD.

Who invented DVD?

No single person or company can claim invention of the DVD and indeed it took time to agree on a single standard. A consortium of ten companies – Hitachi, JVC, Matsushita, Mitsubishi, Philips, Pioneer, Sony, Thomson, Time Warner and Toshiba – was responsible for developing the official specification, with contribution from many other firms. Co-ordination is now in the hands of the DVD Forum (information at http://www.dvdforum.com).

Traps for the unwary

The term 'wide screen' is vague so it's worth distinguishing anamorphic encoding from 'letterbox' presentation. Some discs, like Buena Vista's *The Rocketeer*, have the dual layer set up so that the film can be watched on a 4:3 television in letterbox (first layer) or on a 16:9 wide screen television in anamorphic (second layer).

If owners of 4:3 television try to watch the anamorphic version they will see a squashed image on the screen. Some companies have cheated by offering films only in 4:3 letterbox, which is fine on a 4:3 screen, but markedly inferior when 'blown up' to fit a wide screen set. Purchasers may be unable to tell the difference when buying a disc.

With few wide screen sets in use, the problem has not been highlighted yet. But anamorphic versions are superior in picture quality, with the wide screen set unstretching the image to provide a high-quality wide screen effect.

'Flippers' – techno-babble for sandwich discs – are causing concern to some users, who fear that the sandwich discs won't last. There is a good chance these discs may warp in storage under hot conditions.

A problem with dual layer discs is that there is always a delay when switching from one layer to the other. Some films have sequenced this on a fade-to-black to minimise the effects of the delay for the viewer.

decode microscopic pits embedded in the disc along a spiral track, into audio, video or computer data information. The similarity with CD ends here on account of the higher storage capacity of DVD, which is from 4.7Gbyte to 17Gbyte, with access rates of 600kbit/s to 1.3Mbit/s.

The microscopic pits are in fact reduced in size, from 0.83μm to 0.40μm, and placed closer together. The gap is reduced from 1.6 to 0.74μm. Since the standard CD laser could not read this tightly packed information, a new laser, with a thinner beam and shorter wavelength of 640nm instead of 780nm is employed.

Video data is compressed for storage in MPEG-2 format and on replay is read off the disc at a constant speed of 11.08Mbit/s into a buffer. Conversion to analogue video data follows and combined with audio information for playing out to the viewer's audiovisual system, **Fig. 3**.

Capacity can be increased still further by constructing a disc with two layers of information and using a dual-focus laser to read information from one layer only. This gives the ability to store 8.5Gbyte on one side of a disc.

It is also possible to glue two discs back to back, creating a dual-sided sandwich or 'flipper' disc with a total capacity of up to 9.4Gbyte, or 17Gbyte using dual-layer discs. A good number of movies already take advantage of this capability to include both standard and wide screen versions on the same disc.

Divergent and incompatible audio encoding problems are now history. Generally, DVD software uses either Dolby Digital for USA and some European discs, or MPEG 2.0 or 5.1 sound for European discs.

Some low-end players do not have the Dolby Digital decoders and users must rely on the MPEG stereo sound tracks. All PAL DVD players can play Dolby Digital sound, but NTSC players do not play MPEG audio tracks.

Multiple standards

A common misconception holds that DVD has created a universal television playback system. But this is not the case. It suffers the same NTSC versus PAL problem as videotape and laserdisc.

The digital MPEG video information stored on DVD is formatted for one of the world's two main television systems, NTSC (525/60) or PAL/SECAM (625/50). Differences between discs intended for playback on these different systems include picture size and pixel aspect ratio (720 by 480 *versus* 720 by 576), display frame rate (29.97 *versus* 25), and surround audio. You will find more on this at the 'UK DVD FAQ' web site.

There is also variation in the degree to which players can handle 'alien' discs, although all DVD players sold in PAL countries play both PAL and NTSC – as far as regional coding schemes permit.

Some European machines can also output a PAL-encoded version of American NTSC discs, albeit with slight motion judder, and this can be recorded on tape.

Consumer choice

High technology ought to command

DVD and CD compared

Parameter	CD	DVD
Disc diameter	120mm	120mm
Disc thickness	1.2mm	1.2mm
Disc structure	Single substrate	Two bonded 0.6 mm substrates
Laser wavelength	780nm (infrared)	650 and 635nm (red)
Numerical aperture	0.45	0.60
Track pitch	1.6μm	0.74μm
Shortest pit/land length	0.83μm	0.4μm
Data layers	1	1 or 2
Data capacity	Approx. 680Mbyte	Single layer: 4.7Gbyte x 2 Dual layer: 8.5Gbyte x 2
Reference user data rate	Mode 1: 153.6kbyte/s Mode 2: 176.4kbyte/s	1108kbyte/s, nom.

Courtesy Sony Corporation

DVD-R and CD-R compared

Parameter	DVD-R	CD-R
Media Type	Write-once	Write-once
Wavelength (Recording)	635-645nm	775-795nm
Wavelength (Reading)	635-650nm	770-830nm
Recording Power	4-11mW	4-8mW
Numerical Aperture (Recording)	0.60	0.50
Numerical Aperture (Reading)	0.60	0.45
Reflectivity	R14H>0.5	RTOP>0.65

Courtesy Pioneer

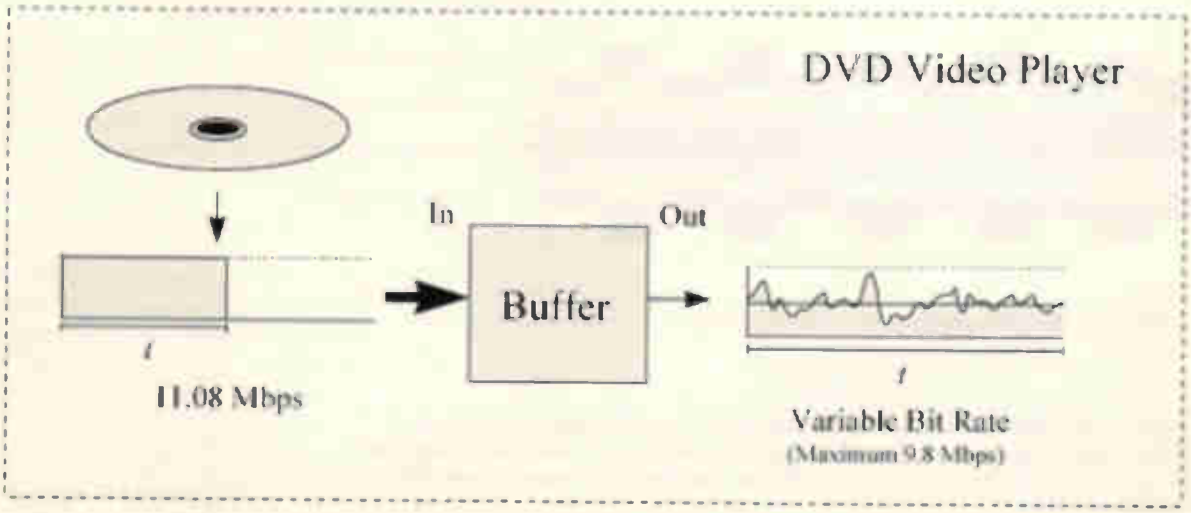


Fig. 3. Translating digital data to analogue video.

high prices but players and discs both cost less than might be expected. In the UK, deck prices start at around £300. While they rise to nearly £9000, there's a broad selection of machines in the sub-£500 arena.

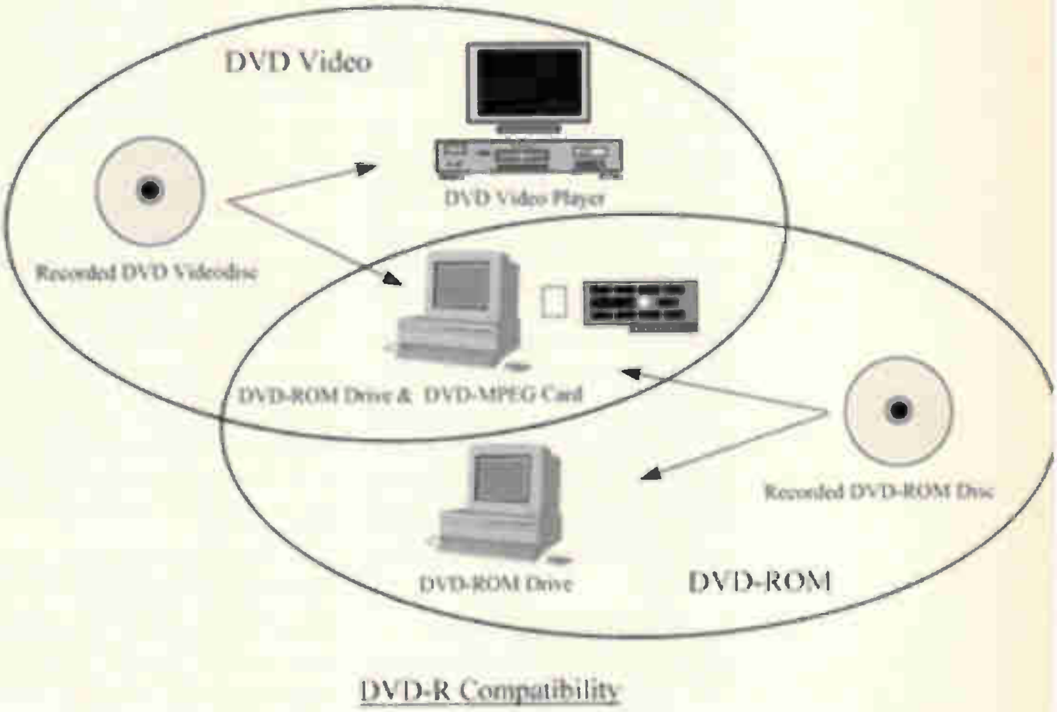
These are all stand-alone decks for use with TV or monitor screens or in home cinema systems but 'bare' DVD drives can also be had at prices under £100 for installing in personal computers equipped with sound and video cards. Many would argue though that watching movies on a PC is a less than satisfying experience and the alternative of hooking up a PC to a TV or home entertainment centre is not the height of convenience.

The range of DVD titles is mainly top-selling movies plus a sprinkling of documentaries, adult titles, sports performances and children's cartoons. In Britain around 1000 titles are available, priced between £9.99 and £19.99.

Many of the movies include additional material such as cinema trailers, cut scenes and production notes, and the list is growing constantly. At least double the number of titles has been released in the USA, but there can be problems replaying them outside the US on account of regional coding, as discussed below.

Compatibility corner

DVD-ROM players can play old CD-ROMs, CD-I discs, and video CDs, as well as new DVD-ROMs. Newer DVD players – so-called second-generation DVD or DVD-2 – can read CD-R and CD-RW discs. DVD-RAM drives can also act as DVD-ROM and CD-ROM players.



Compatibility between DVD's video and data variants.

Paradise lost

Earlier on, I hinted that not all was sweetness and light with DVD – at least as far as consumers were concerned. The twin sources of this aggravation are encryption and regionalisation.

Encryption is the simpler concept to understand. Low-cost CD writers and blank discs mean that making illegal copies of CD-ROMs is child's play. It's also completely uncontrollable today

To prevent a repeat performance of the CD copying problem, DVDs are encrypted via sophisticated security encoded into them. The name of this security system is Content Scrambling System. This system prevents media files being read directly from the disc without a 5-byte (40-bit) decryption key. Copying is thus prevented – in theory.

Although DVD's security system was supposed to be hacker-proof, it turned out not to be idiot-proof. An error by a software licensee allowed some gifted programmers in Norway to write a 60 kilobyte piece of code – named DeCSS – that allows DVD films to be saved to hard disk and hence copied with the file encryption removed.

The threat to film producers is largely hypothetical at present: recordable DVD has at best 2.5Gbyte capacity currently – or 5.2Gbyte for double-sided discs. Given that DVD movies can range from 4.7 to 9.4Gbyte, this means that direct DVD copying is not feasible.

High-capacity recordable DVD drives are expected during 2000 however. These will have the 4.7Gbyte capacity needed for these illegal pursuits – unless the copy protection issue is settled by then.

DVDs can of course be copied to tape, which is why some distributors have applied Macrovision anti-copy protection to the video signal – for example Rank's Region 1 DVDs. This works very effectively, and is better than the VHS version. It can be subverted with the usual Macrovision decoder boxes though.

Regionalisation explained

The desire of Hollywood movie producers to control the release of their products in different world markets led to a scheme to restrict playback called Regional Coding.

DVD players and discs are coded according to the world region of their intended use. As a result North American (Region 1) discs will not play in a European (Region 2) player – in theory. The conflict between the 'rights' of producers and consumers has been hotly debated but the desire for 'code-free' machines that can play discs from any part of the world means that these are not hard to find.

Many dealers will provide modified machines on request. Even if they don't, there are many advertisers on the WWW who will supply the necessary EPROMs or microprocessors for users to upgrade their own machines.

Regional round-up

These are the region codes for DVD:
Region 1 – North America and Canada
Region 2 – Europe and Japan
Region 3 – Asian Pacific
Region 4 – South America/Mexico/Australia/New Zealand
Region 5 – Africa and the Eastern Europe, Russia
Region 6 – China and Hong Kong

Region 0 (universal) discs can play on any machine and some of these discs are already available. Most of these are titles such as IMAX Travelogues and short documentaries. DVD drives for PCs are supplied uncommitted to a particular region. The owner then sets the desired regional affiliation in software, sometimes without the ability to change this subsequently.

See the section headed '**Resources**' below.

Adapting or 'chipping' a Region 2 player to replay Region 1 discs costs between £20 and £100. Bear in mind though that such modifications of course invalidate the manufacturer's guarantee.

Hardware changes are not always necessary: some machines require just a simple three-digit numerical code typed in sequence on the remote control to prime the machine to accept NTSC discs.

The wider choice of DVD movie titles available in the USA led to a healthy business importing them to Europe. But this soon came to an end when discs were seized on account of not having UK censor certification.

It is illegal to sell uncertified imported DVDs commercially in Britain. Companies attempting to side-step the rules by renting them instead hit a similar barrier.

There is nothing to prevent individuals from importing discs for private use, by mail order Internet purchase or on a personal basis. It is certainly not illegal to own them, so long as they do not offend any decency considerations of course!

Fast forward to the future

It doesn't take a genius to realise that DVD hardware and software prices will fall further, increasing market penetration at the same time. The biggest imponderable is the extent to which DVD could displace VHS recorders – or at least relegate them to a matter-of-fact device for simple time-shifting.

Early adopters have already shown the way; they have demonstrated that the DVD's enhanced visual and aural quality, and its greater flexibility of use, is paramount for them.

There are also those who argue that the trivialisation of broadcast television today – 200 channels with nothing on – will focus ever more attention on quality entertainment that's pre-recorded. Currently top-selling releases are issued on DVD as well as VHS but the time must eventually come when software houses drop the VHS format, even if this is not for another decade or so.

It's notable that an increasing number of art movies and special 'connoisseur' packages of archive material are being released on DVD alone; this is not only because higher prices can be charged, but also because serious viewers are no longer satisfied with the quality – and vulnerability – of VHS tape. BBC Enterprises has hinted, for example, that future releases of *Doctor Who* and classic drama and concert performances will appear only on DVD.

In the data arena many experts believe that DVD-ROMs will eventually replace CD-ROMs, possibly sooner rather than later. The greater data storage capacity means that graphics can be of photo quality, rather than animated cartoon-style graphics.

DVD-based in-car navigation systems could contain street plans and road atlas maps of all Europe on a single DVD-ROM. Alpine plans to offer such a system during 2000.

Your mileage may vary

Understandably, opinions vary on how fast all this will come to pass. Blockbuster Video – the self-styled global leader in rentable home entertainment – has committed to introducing DVD in nearly 5000 of its stores internationally in what it terms the fastest growing packaged media playback device in history. The company aims to be the leading global source for rented DVDs – a product which its analysts predict would be in more than four million American households by the end of 1999 and in 40 to 50 million households there by the year 2007.

1 in 100 UK homes watches DVD movies

Others take a more sanguine approach. Speaking at 'DVD Forum Conference 99', Jim Bottoms of British consultants Understanding & Solutions noted that while DVD movies had achieved a 5 per cent penetration of US homes, fewer than one per cent of homes in Europe had adopted DVD.

Part of this is due to the difference in the availability of titles – 5500 in the United States versus a mere 600 in Europe.

Jim's prediction was that DVD players would be found in 10 per cent of US homes by the end of 2000, increasing to 25 per cent by 2002. Europe on the other hand would lag, barely topping 1 per cent in 2000 to reach 5 per cent by 2003.

DVD-ROM software for PCs, both applications and games, would trail further behind, he declared, although this contrasts strangely with Dataquest's forecast of 25 million DVD drives being installed in PCs during 1999, compared with around 3 to 4 million home-DVD video units.

Glittering future?

Part of the uncertainty over DVD's future relates to the importance that viewers attach to actually owning their movies.

Currently you need to physically have and hold a tape or disc to replay it, but great play is made of video-on-demand in future. Whether it is delivered by satellite, digital cable TV or by DSL down telephone lines is irrelevant; if people have access to hundreds or thousands of titles on demand at an acceptable cost there would be no reason to invest in DVD.

Of course this is a mighty big 'if' but it could alter the way that casual viewers obtain their entertainment. Time alone will tell. ■

Many thanks to Andrew Henderson for technical assistance in preparing this article.

Resources on the Web

http://www.videodiscovery.com/vdyweb/dvd/dvdfaq.html#1.19
The so-called **Official DVD FAQ** and certainly one of the most comprehensive.

http://movieuk.com/dvdfaq.htm
The **UK DVD FAQ**, with a lot of informative material including machine-by machine hacks.

http://www.btinternet.com/~bryan.welsh/dvdfaq.htm

http://home.clara.net/laserdisc/DVD-ROM.htm
DVD & Laserdisc UK Website, with various DVD-ROM 'region' hacks and downloadable software.

http://www.activewin.com/drivers/dvd_rom_drivers.shtml
Device Drivers, Files and Firmware Downloads for DV-ROM systems.

http://www.dvdclub.com/message/
DVD Club bulletin board, for people sharing all manner of hacks and cracks.

http://www.cdrom-guide.com/dvdforum.htm
DVD Discussion Forum.

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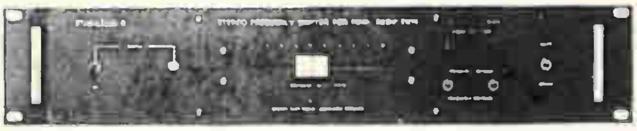

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ECC88	5.00	EL509 519	13.00	7581A	11.00	4 Pin	3.30
ECF82	5.00	E84L/7189A	6.50	807	9.00	4 Pin	5.00
ECL82	5.00	KT66	9.50	811A	11.00	4 Pin	11.00
ECL86	5.00	KT77	12.00	812A	34.00	4 Pin	
EF86	5.50	KT88	12.50	845	30.00		15.00
E80F	10.00	KT88	21.00	**RECTIFIER TUBES**		5 Pin	3.00
E81CC	6.80	KT88	60.00			7 Pin	4.50
E82 CC	8.00	PL509 519	9.00	EZ80	4.00	9 Pin	5.00
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CIRCLE NO.112 ON REPLY CARD

Analyse shutter speed

Ever wondered how precise and repeatable your camera's shutter is? Douglas Clarkson explains how to check a camera's shutter using a dual-channel light-pulse timing system.

There are many situations in photography where it is desirable to be able to evaluate how well a piece of equipment is working. A good example is the performance of a mechanical shutter. When examining shutter functioning, it is also useful to be able to verify any associated flash units and their shutter synchronisation.

Table 1. Range of count values relevant for a four-digit display derived from a range of clock frequencies.

Exposure (s)	Count 1MHz	Count 100kHz	Count 10kHz	Count 1kHz	Count 100Hz
1/5000	200	20	2	-	-
1/2000	500	50	5	-	-
1/1000	1000	100	10	1	-
1/500	2000	200	20	2	-
1/250	4000	400	40	4	-
1/125	8000	800	80	8	-
1/60	-	1667	167	17	1
1/30	-	3333	333	33	3
1/15	-	6667	667	67	6
1/8	-	-	1250	125	12
1/4	-	-	2500	250	25
1/2	-	-	5000	500	50
1	-	-	-	1000	100
2	-	-	-	2000	200

At first sight, examining a shutter seems a simple task. On delving further though you will find that it throws up some surprising complexities. It is important to keep things in perspective though when assessing how the subtler aspects of shutter performance affect picture quality.

In terms of assessment of quality of camera performance, there should be little shot-to-shot variability of timing. Additionally, the indicated exposure time should correspond with that selected.

This article describes a simple electronic device that I have developed and built to measure the exposure time of manual camera shutters. It also facilitates the checking of flash/shutter co-ordination as an integral part of camera function.

Timing method

One of the problems of dealing with camera systems is the wide variation of exposure times that may need to be measured.

I found that the best way of measuring exposure times is to use digital techniques. I created an enabling window signal that was triggered on by a particular light level detected on shutter opening, and off as the light level falls due to the shutter closing. While this window signal is 'open', clock pulses are gated 'on' and counted, recorded and displayed.

At the shortest exposures, a fast counting clock is required. By using a 1MHz master clock with derived clocks of 100kHz, 10kHz 1kHz and 100Hz, a four-digit counter device can be configured to count over a wide range of exposure times, indicated in **Table 1**.

General design considerations

For the counting and displaying part of the design, I had to choose between devices based on light-emitting diodes and those incorporating liquid-crystal digits. LED-based devices are simpler but consume significantly more current. In this design, I chose LEDs for simplicity.

Use is made of the Harris ICM7217IJI common-anode counter and display module. It provides 4 digits of resolution. Separate channels of light signal detection A and B are structured as in **Fig. 1**.

Overrange signal for count display is indicated as a status LED. Control

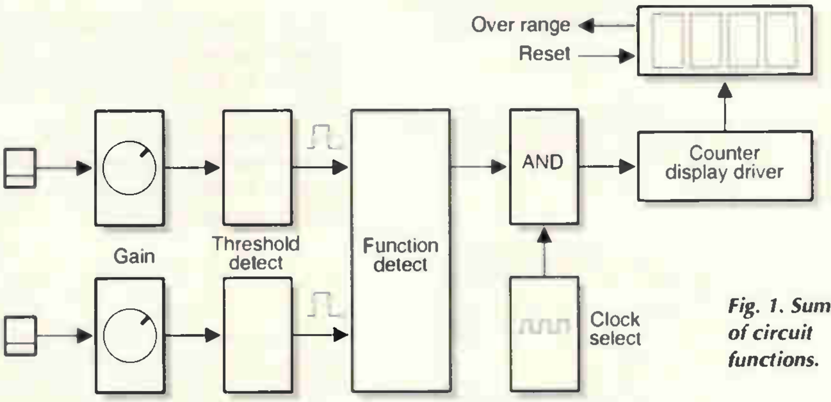


Fig. 1. Summary of circuit functions.

options are as indicated in **Table 2**. Clock rate can be selected to optimise count resolution as indicated in Table 1. The unit is powered from a 9V power unit since the LED display requires significant current drive.

The sensing element of the circuit is a high-speed silicon photodiode. A typical voltage signal from a photodiode circuit placed in the rear of the camera would be as indicated in **Fig. 2**. This indicates the rising edge AB, the plateau BC and the falling edge CD.

The response speed of the photodiode and the associated electronics can be assessed as of the order of a few microseconds and not to contribute significantly to measurement errors.

Timing considerations

Next I had to decide which light thresholds are appropriate for triggering and terminating the train of clock pulses? Consistency of approach is important here. If a consistent level relative to the plateau level can be selected, then this helps to provide that consistency.

With a light source shining through the camera and with the camera timing bulb B option active (shutter open), the sensitivity is adjusted until the measured signal equals a reference level established separately in the circuit of 2.5V. This represents 100% of the expected 'plateau' voltage of the light detector for its particular location on the back plane of the camera.

I made measurements and selected 10% of this value as the threshold value for activating/deactivating the counter when exposures are being measured. This provides a means of establishing reproducible threshold limits for exposure measurement as indicated in **Fig. 3**.

Up to five separate clock rates can be selected – 1MHz, 100kHz, 10kHz and 1kHz and 100Hz. The clock signal is derived from a 1MHz quartz crystal and divided down by four separate divide-by-10 counters.

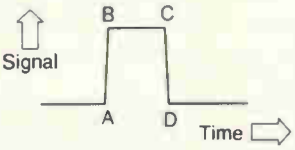


Fig. 2. Typical signal derived from light sensor circuit on back plane of SLR camera – rapid rise of signal AB to plateau BC followed by rapid fall off CD.

A reset button is available to zero the indicated count value. A separate buffered output of the photodiode signal is available for monitoring on an oscilloscope. Also, for calibration purposes, a separate buffered output is present for the 1MHz clock signal. Basic accuracy of the quartz crystal is typically 100 parts per million or 0.01%.

Initial results and observations

Initial results are presented in **Table 3** for a Miranda 35mm SLR camera.

Exposure times are essentially correct up to 1/500s. Exposure time measurements of 1/1000s and 1/2000s can be achieved by adjusting the size of the width of the light detector. As narrow a slit detector as possible should be used.

Table 3. Initial results for Miranda 35mm SLR.

Set time (s)	Measured time (s)	Expected time (s)	error (%)	top/base error (%)
1.0	0.9752	1.00	–2.48	
0.5	0.4513	0.5	–9.74	
0.25	0.227	0.25	–9.2	
1/8	0.1258	0.125	+0.64	
1/15	0.0627	0.0667	–6.0	
1/30	0.0308	0.0333	–7.5	
1/60	0.0174	0.0167	+4.4	2.4
1/125	0.01006	0.008	+25.8	4.8
1/250	0.00446	0.004	+11.5	9.5
1/500	0.00235	0.002	+17.5	18.3
1/1000	0.001825	0.001	+82.5	28.5
1/2000	0.00094	0.0005	+ 88.0	40.6

Table 2. Main control options.

Option	Comment
A only	pulse duration channel A only
B only	pulse duration channel B only
A + Not B	transit time of shutter
B + Not A	transit time of shutter

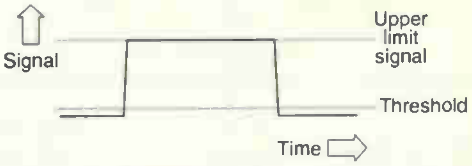


Fig. 3. Derivation of effective threshold levels for timing measurements. A plateau level signal is derived and a threshold level set at 10% of this for the start and stop of the timing interval.

Table 4. Data from Cosmic Symbol leaf shutter camera.

Set time	Measured time	Expected time	% error
1/15	0.06165	0.06667	–7.5
1/30	0.03095	0.03333	–7.2
1/60	0.01546	0.01667	–7.3
1/125	0.00882	0.008	10.25
1/250	0.00455	0.004	13.75

It can be shown that for the model of camera, exposure times were shorter towards the top of the film and that these differences were more evident for shorter exposure times. Probably, however, these effects would not be noticed in photographic work.

Table 4 is results from an old Cosmic Symbol leaf shutter camera available in the sixties. Considering the camera's age, its shutter performance is rather good. The size of light detector is not as critical for this type of application.

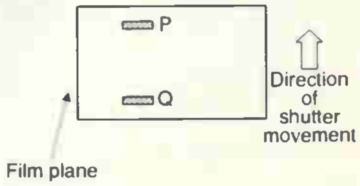


Fig. 4. Separate detectors P and Q at extremes of travel of focal plane shutter allow time of transit of shutters to be directly measured.

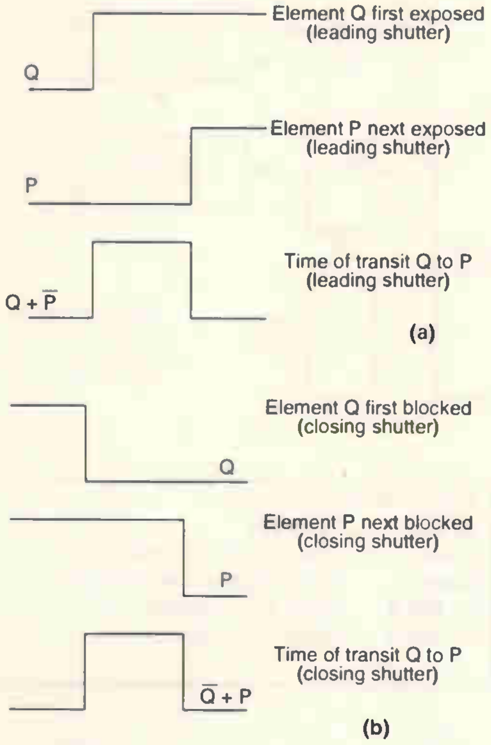

Element Q first exposed (leading shutter)
Q
Element P next exposed (leading shutter)
P
Time of transit Q to P (leading shutter)
$Q + \overline{P}$
(a)
Element Q first blocked (closing shutter)
Q
Element P next blocked (closing shutter)
P
Time of transit Q to P (closing shutter)
$\overline{Q} + P$
(b)

Fig. 5a). Logic signals to derive transit time of leading edge of leading shutter. An AND gate is used to add Q and the inverse of P.
5b). Logic signals used to derive transit time of leading edge of closing shutter. An AND gate is used to add P and the inverse of Q.

Using a small circular sensor, it was possible to detect variations in exposure times as the sensor was moved across the axis of the aperture.

Benefits of dual light sensors

It is relevant to measure the 'on' duration of the camera shutter, but there are also other considerations that are relevant.

Consider that in **Fig. 4**, two sections P and Q are selected, together with an exposure during which the entire back plane is exposed. Next, assuming the first lead edge rises up from top to bottom then signal Q will register before that of P.

On closing the shutter, and assuming the leading closing edge rises vertically, Q will be closed off before P.

The associated electronic signals are indicated in **Fig. 5a)** for the leading shutter action and **Fig. 5b)** for trailing shutter action.

Note that for the opening shutter, if the signal from P is inverted and combined in AND logic, the resulting Q + Inverted P gives the time during which one detector receives a signal while the other does not.

The mean time of transit of the leading shutter was measured as 6.3ms and that of the trailing shutter as 6.1ms – equivalent to a mean velocity of 387cm/s across the 2.4 cm film width aperture.

For either the leading or the closing shutter, the speed between any two vertical lines can be measured using this method. The minimum separation of detectors for such measurements is around 1cm. Again this technique depends on the use of a very narrow detector slit.

Shutter types

Leaf shutters operate on the principle of the iris. They comprise a number of interconnecting curved blades that initially expand to open the aperture and then collapse back to close it.

Capable of shutter speeds of up to about 1/500th of a second, this type of shutter has two or more leaves and is found in 35mm compact cameras.

Leaf shutters have an inherent drawback in that the iris, on opening, causes the centre of the film to be exposed to the picture for longer than the periphery.

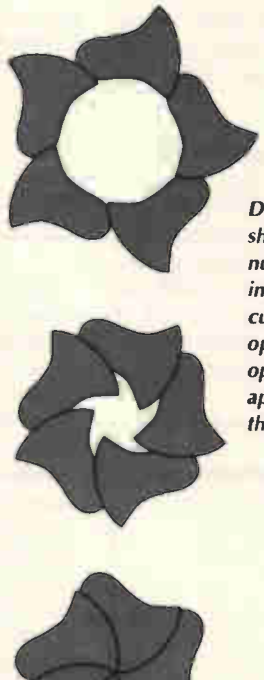
When the shutter is fully open, light from a flash can reach the whole of the exposed film. This makes it important to have good synchronisation between the open aperture and the flash output.

Focal-plane shutter

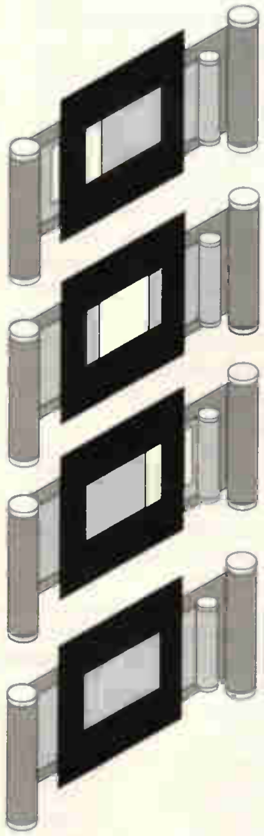
The focal plane shutter is a key element of single-lens reflex camera technology. Such shutters are either horizontal or vertical types – with vertical types being typically faster.

Such shutters rely on a principle similar to that of a roller blind, where there is a leading edge to open the shutter and a separate trailing edge to close it. In long exposures, for example 1/4s, the main element of timing relates to the time that the film is totally exposed.

At increasingly shorter exposures, a successively narrower strip of film is exposed. The limit of speed of exposure is typically around 1/2000s, though this is intrinsically dependent on the speed of the shutter movement.



Detail of leaf shutter: A number of interconnected curved blades operate to open the aperture and then close it.



Detail of focal plane shutter: The focal plane shutter can be considered to comprise two independent blinds that move across the focal plane where the film is present. At slow speeds the entire film aperture is exposed while at successively shorter exposures, the blinds move to effectively direct a narrowing slit of exposure across the film.

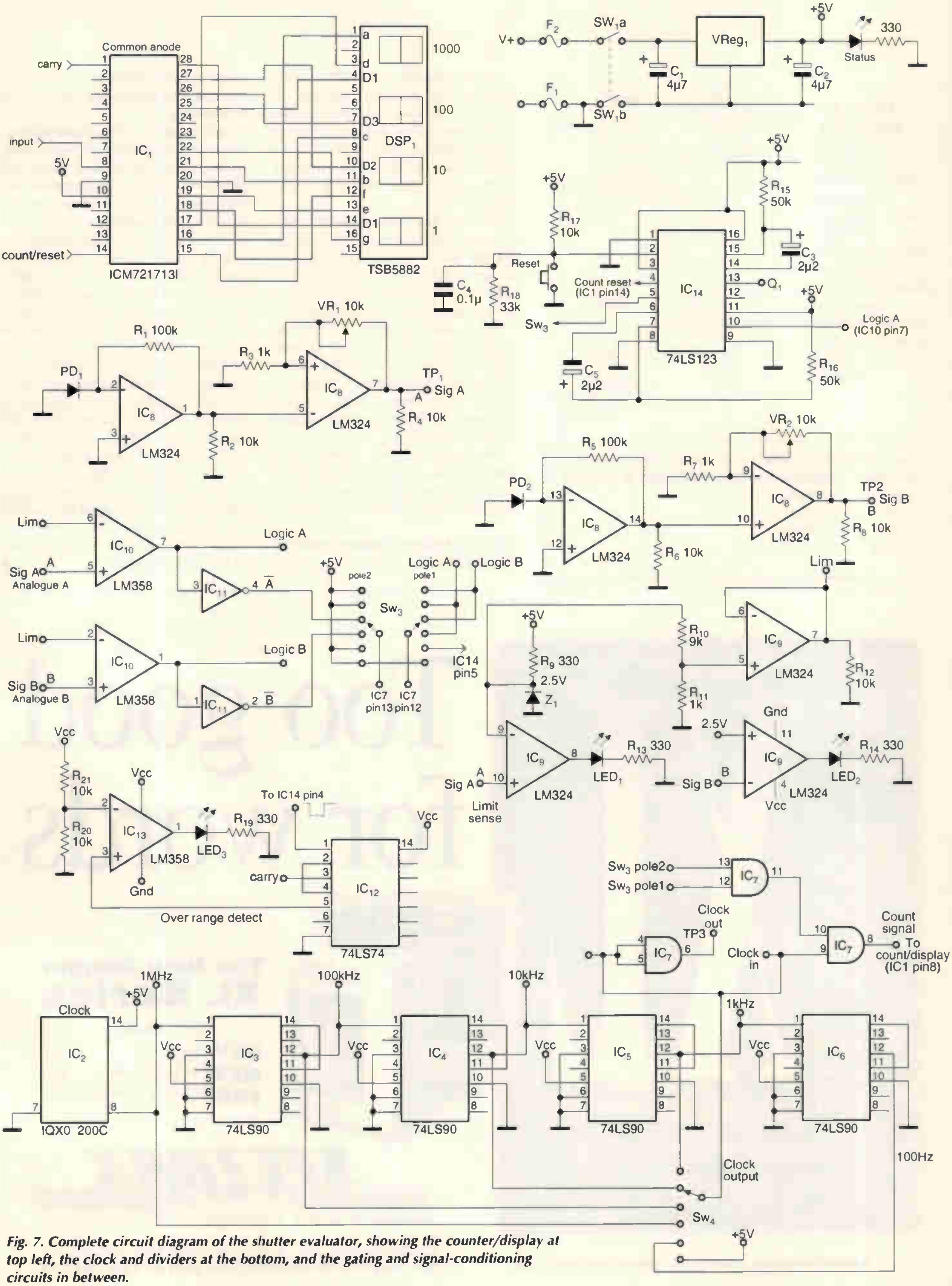


Fig. 7. Complete circuit diagram of the shutter evaluator, showing the counter/display at top left, the clock and dividers at the bottom, and the gating and signal-conditioning circuits in between.

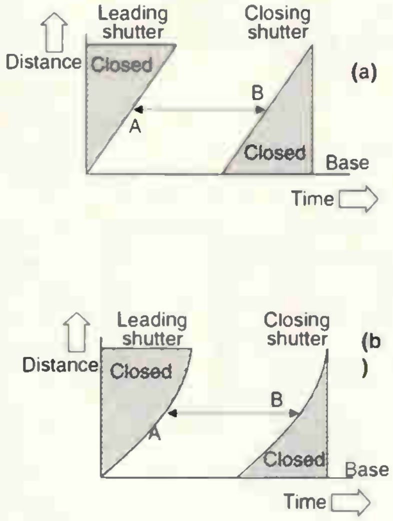


Fig. 6a). The 'perfect' focal plane shutter where velocity profile of leading and closing shutter edges are identical. 6b). Example of a notional focal plane shutter with imperfectly matched velocity profiles of leading and closing shutter.

Table 5: Estimated exposed 'slots' for the Miranda camera.

Set time	Estimated exposed slit width @ 387cm/s
1/250s	1.55cm
1/500s	0.77cm
1/1000s	0.39cm
1/2000s	0.19cm

For the example of a 1/2000 second exposure at a mean velocity of 387cm/s, a strip of approximate width 2mm will be exposed. You can see the 2mm slit size as being some kind of basic dimensional limitation to the size of film field exposed in this way.

One essential requirement for measurements at such exposure times is that the size of the detector should be as small (slit like) as possible in order to give accurate values – ideally less than 1mm in cross section.

Initial measurements of exposure times below 1/250 second showed large errors due to use of a sensor with a large detecting area. Estimated values of estimated strip size are indicated in **Table** 5 for an average velocity of 387cm/s.

Obviously higher performance cameras can cope with significantly smaller shutter times by way of having much faster shutter movements.

The focal-plane shutter revisited

The focal plane shutter is an artful solution to the problem of consistent, uniform exposures over the 35mm film surface.

Such a shutter is the combination of two separate mechanical components, defined here as the leading shutter and the closing shutter. By measuring the time taken to cover different portions of travel in the backplane of each shutter, such as described previously, the leading edge shutter's velocity is found to increase as it travels upwards. This is also true for the leading edge of the closing shutter as it rises upwards.

Provided, however, the relevant velocity profiles are essentially matched, this will lead to uniformity of exposure across the film. Such effects are more significant for exposures of less than 1/125 of a second and during which only a portion of the film is exposed at any one instant. These effects are shown in **Fig. 6**.

The system can also be used to evaluate flash synchronisation by comparing the time of the flash as measured on the film plane to the actual duration of the flash exposure. ■

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Lundahl's 1660 valve interstage transformer

Hybrid audio power

Hybrid valve/fet amplifiers usually need some form of DC blocking. Traditionally, transformers have been frowned on because of their non-linearities, but what happens when you use a modern high-performance transformer? Wim de Haan outlines his findings.

Performance measurements

–1dB frequency response, 1W	14Hz-45kHz
THD + N at 1W:	0.26 %
THD + N at 10 W:	0.36 %
THD + N at 36 W:	0.6 %

Hybrid amplifiers using a tube as the driver stage and power mosfets for current handling are becoming popular. Such amplifiers need some form of coupling.

For convenience, a capacitor is usually used. The quality of this capacitor is important. It is directly in the signal path and has a major effect on the overall sound quality of the amplifier.

For my design, I decided to experiment with a high-performance c-core interstage transformer made by Lundahl, and in my view the experiment was a success.

Hybrid amplifier revisited

This hybrid amplifier uses a 5687 tube, SCR coupling capacitors and Magnatec power mosfets. I find the results extraordinarily good.

Inspiration for the design came from a Japanese magazine from 1995. It described a design with an op-amp, an interstage transformer and two power mosfets.

After reading the article though, a question remained. Transformers had disappeared from modern designs because of their poor performance hadn't they? But then using a transformer would get rid of the coupling capacitor, so I decided to give it a try.

I chose the 5687 and Magnatec mosfets having had success with them in previous designs. The fets are specially made for audio and are reasonably priced.

The BUZ900 has a TO-3 case while the BUZ900P is in TO-247. There is also a version with two n or p-mos fets in parallel in one case.

Interstage transformers from Tango and Tamura are simply too expensive to experiment with. After a period surfing on the Internet I found an more reasonably priced alternative from Lundahl. It is also easily obtainable and has good specifications.

Circuit details

I have built, tested and used the circuit diagram shown here.

On building my prototype, I noticed it oscillated, hence I added an *RC* network. Another solution may be to change the gate resistor from 249Ω to 1.5kΩ.

Unloaded, the secondary winding of the transformer has a peak of 8dB at 30kHz. A 47kΩ resistor in parallel with the secondary winding gives the best compromise between distortion and damping/frequency response.

With this resistor in place, the amplifier has a peak of a 0.5dB at 20kHz and –1dB points at 14Hz and 45kHz – quite an achievement for 1:1+1 transformer coupling.

An attempt to improve damping of the secondary winding using a parallel *RC* network was unsuccessful.

The transformer described above is a Lundahl 1:1+1 interstage type, namely an LL1660. It is obtainable in three versions – one for push-pull operation, and two for single-ended use. The single-ended types differ in current-handling capability. One handles 18mA, the other 10mA. I chose the latter.

The configuration chosen for the interstage wiring is suitable for single-ended to push-pull use with a primary current of 20mA. The 5687 is compatible with this current.

An alternative is an interstage transformer from Border Patrol. This English company also manufactures a usable 1:1+1 interstage transformer, part number ITPP-860. I have no experience of this product though.

Listening results

Not having built the second amplifier of the stereo pair yet, I cannot give a full report on the sound of the amplifier. In mono though, it sounds promising. Notable points are good audibility and clarity of the sound. The low end is dry, punchy and tight.

I have hesitated building the second channel because I have not yet decided whether to try the design with two n-type mosfets.

Finally, my thanks to Guido Tent whose original work inspired this design ■

Further information

More information on this hybrid amplifier
http://www.homestead.com/whaan/files/index.html

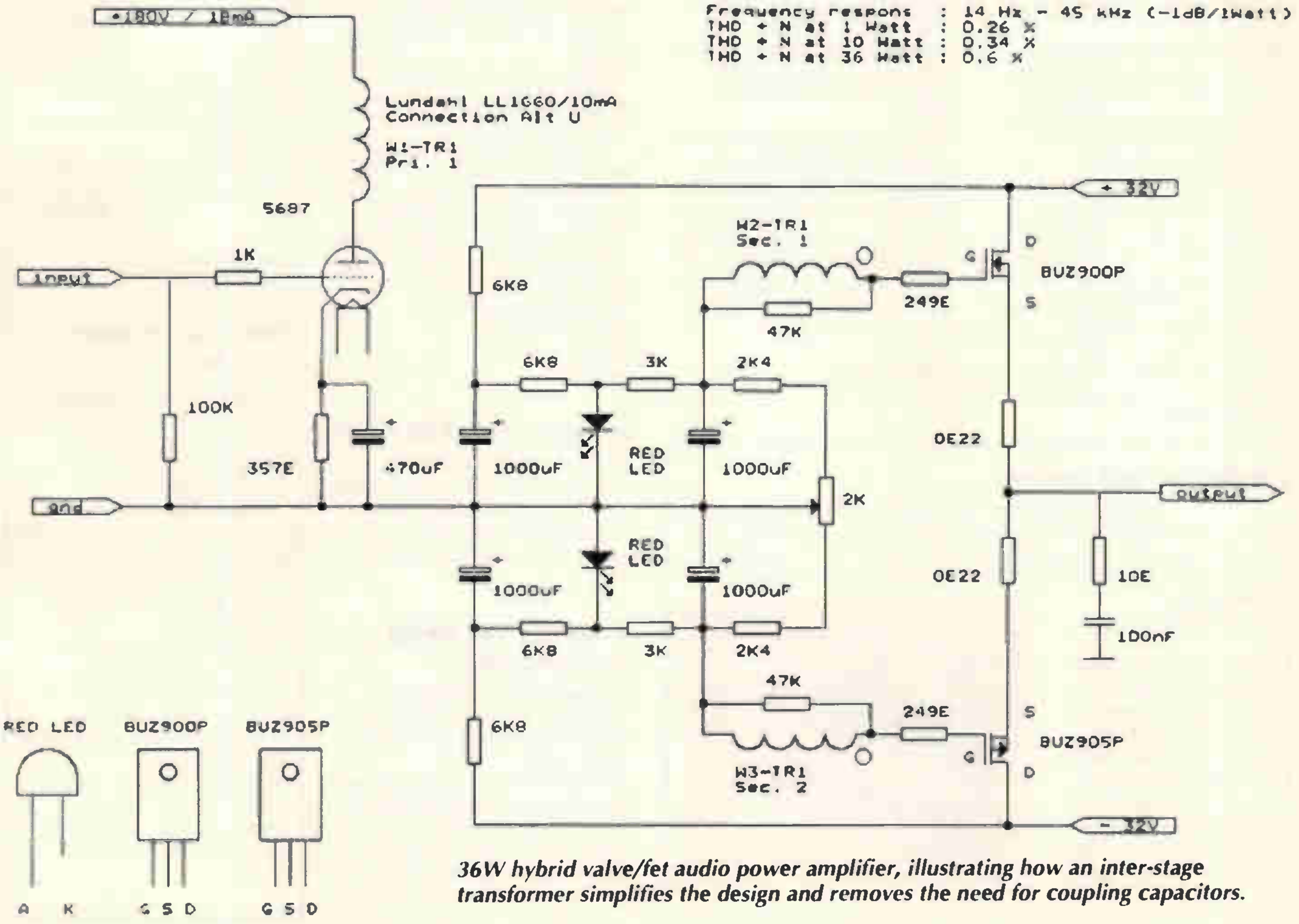
Lovers of classical piano music
Visit: http://www.wdehaan.demon.nl/

Information on Lundahl transformers
Datasheets, Internet : http://www.lundahl.se/
Telephone : + 46 176 13930

Information on Magnatec power mosfets (datasheets)
Internet : http://www.semelab.co.uk/magnatec.htm
Telephone : + 44 1455 554711

Border Patrol transformers
Telephone: + 44 1273 276716

Lundahl Belgium
Internet : http://www.degryse.com/
Telephone: + 32 3 281 32 44



36W hybrid valve/fet audio power amplifier, illustrating how an inter-stage transformer simplifies the design and removes the need for coupling capacitors.

LETTERS

**Letters to "Electronics World" Quadrant House,
The Quadrant, Sutton, Surrey, SM2 5AS
e-mail jackie.lowe@rbi.co.uk using the subject heading 'Letters'.**

Society's scientists

Richard Wilson's leader in the November issue, about the loss of public confidence in science and scientists, could have added another couple of points. Perhaps the most significant is that the events most destructive of public confidence have been in matters in which there was a large commercial or political interest.

In old-fashioned terms, power and greed corrupt, and the powerful and greedy will always try to claim the support of 'eminent scientists' if they think it useful – though usually without remarking that those eminencies are financially dependent on their employers.

This also influences the traditional scientific process. Despite what Dr Peter Cochrane is quoted as saying, it is often very difficult to repeat experimental results in the early stages of discovery when concepts are tentative and the significant parameters poorly understood.

High-energy experiments are a good example, where results are accepted as evidence that in other fields might be dismissed as nebulous or statistically inadequate, simply because it is so hard to get results at all. When there are also others working in the field, who are eager to get first credit – recall the undignified scrambling and lobbying attendant on Nobel Prizes – or whose employers want royalties on potential inventions, it is not surprising that some scientists publish prematurely.

The very publication process is distorted by the editors of supposedly respectable journals, who will not accept papers written in simple comprehensible language and format but insist on a dreadful artificial dialect that obscures meaning. Popular newspapers can scarcely be blamed if they misunderstand such jargon.

Some of the journal editors are also working scientists who use their editorial positions to suppress work that does not suit their own positions; and the peer review process, though laudable in its original purpose, can be needlessly destructive.

There are too many well-known examples of this for it to be worth documenting in a letter. This too leads to original work being published with non-traditional timing or in non-traditional places.

It is clear enough if you read the literature that in science, just as in politics and religion, people of a dogmatic temperament can feel quite at home. There are many anomalies in physics, but dogmatic theorists reject or ignore them (who will work on Faraday's disk?).

If we are to see a return of public admiration, perhaps even trust, scientists are going to have to resist pressures to agree with their employers; need to start writing in words that can be understood; and must actively do something about their dogmatists. That is asking rather a lot, and I am not optimistic.

Roderick Rees
Woodville, WA
USA

Relief for pot famine

I have about 12, dual 2kΩ, tent-turn potentiometers that I can extract from ex-BT pager transmitters. They are manufactured by Bourns. I will remove, test and post one to any reader sending me £5.50 – while stocks last of course. Overseas readers please send an extra £2.00 postage.

John Denby
Huddersfield

Please send your request to 'Pot relief', Electronics World Editorial, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. John sent me a sample, shown in the photo, which, considering the list price is over £20 each in ten-offs, looks like a good buy. Ed.

Post Y2K

Just heard on TV the frenetic news that a trillion dollars has been spent worldwide on hopefully forestalling Y2K problems.

In most local and international cases, utilities, services, communications, police, suppliers, emergency groups, transport, hospitals... are on red alert. More personnel than ever before are simultaneously without any of the season's traditional holidays.

Hype, suspicion and ballyhoo infest computer software systems that are claimed to be poorly dated. The excuse is that these are maybe 20 years old.

Now you can't tell me the entire world is virtually run by such antiques! Nor that a mere 5 to 10 years ago there was no confidence in equipment reaching AD 2000.

Computers are meant to save us time and money. But savings are being monumentally lost as millions of folk are consumed by trivia.

Whoever's fault it is stops nobody buying such rubbish computer systems.

The real problem for most people is their gross overdependence on public resources for non-sustainable home life-support systems. As for ourselves, come what may or may not, we have no worries

Dr Patrick Howden
BackYard Tech
Macleay Island
Australia

Is your VCR Y2K OK?

Many video recorders have calendar settings. If yours gets the days of the week wrong now that it's year 2000, don't throw it away. Simply set the year to 1972.

Comfort for timid components

I wonder if M. Vanthomme's 2N3055s (November 1999 issue) all try to hide at the back of his box of spares?

I suspect there is a rule of thumb that states that it is impossible to make a switched-mode power supply without inductance. As each switch connects a voltage-stiff branch into a current-stiff loop, or vice-versa, a capacitor must necessarily be connected to an inductor.

Aha! what about switched capacitor circuits I hear you say? Well, no, they aren't strictly switched mode (with one bound he was free!).

M. Vanthomme's circuit is no more efficient than a linear one because he is merely conducting current between two voltages. And brutalising his poor 2N3055s to boot. ■

Richard Aston.
Via e-mail

A slip of the Tong

Thank you for publishing my letter concerning alternatives to thermistors in the February issue. Unfortunately a bug intervened and changed my name to Tong.

Mike Cox
Twickenham

Lead-acid comments

It was nice to see my lead-acid battery charger circuit in print in the January issue; it would have been even nicer had you printed it correctly.

Resistor R_{11} is 15mΩ, not 15Ω. Diode D_{10} is drawn as a zener; it is actually a Schottky device.

Ian Benton
Via e-mail

Power-factor meter web site

Michael Slifkin's power factor meter, presented in the February issue, was a complex circuit intended for readers experienced in digital logic circuitry. Unfortunately, because of the complexity of the circuitry, and the fact that the illustrations were divided into sections, several errors crept into the circuit diagram.

The authors are setting up a web-site containing corrections and other information that may be of value to those of you wanting to implement the meter. You can find it at http://optics.jct.ac.il/~slifkin/pfm.htm. This site will also contain the URLs for the data sheets of all the ICs in the circuit.

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CIRCLE NO. 117 ON REPLY CARD

Towards a 500MHz scope add-on

Ian Hickman has been experimenting with a sampling technique that was once widely used in very expensive oscilloscopes. It turns out that using the same technique with modern components, useful performance can be obtained at very ordinary prices. This work indicates that with further development, it should be possible to produce an add-on oscilloscope adaptor providing a 500MHz bandwidth. Watch this space...

Back in the 1950s, oscilloscopes struggled to provide a bandwidth of 50MHz – 25MHz being nearer the norm. A notable Tektronix model, with the aid of a special plug-in, managed 85MHz. This was considered remarkable in its day. Strange to relate, then, that in the late 1950s an oscilloscope appeared which boasted the then incredible bandwidth of 2000MHz.

This Hewlett-Packard instrument, I think the model number was HP260 or maybe HP280, was an entirely new breed of device, called a 'sampling scope'. It achieved its remarkable performance by giving up the quest to handle the incoming signal in real time. Instead, it used a very high-speed gate, operated by a very narrow pulse, to sample the signal at intervals. A trigger circuit derived the timing of the sample pulses from a divided-down version of the input.

It was so arranged that as the spot on the cathode ray tube display moved across the screen, successive sample pulses were delayed by a steadily increasing amount on the trigger pulses. Thus a representation of the waveform under investigation was built up from samples taken from non-contiguous succeeding cycles, rather in the manner of a stroboscope.

The sampling rate was typically limited to a maximum of 100000 samples per second. So when examining a 500MHz signal, a sample would be taken from every 5000th cycle.

The technique has been refined over the years by both Hewlett-Packard and Tektronix, leading to instruments with a bandwidth of 14GHz in the 1970s, and more recently, of 50GHz. But like the original models – indeed like all sampling scopes – these were limited to operating on continuous signals. Clearly the technique does not, by its very nature, lend itself to capturing transients, or fast one-off events of any kind.

Sampling oscilloscopes of the type referred to above, sometimes called analogue sampling scopes, are now obsolete. But the technique has been used as the front end of some digital-storage oscilloscopes. The distinction between these and ordinary DSOs is indicated by their description as 'digital sampling oscilloscopes'.

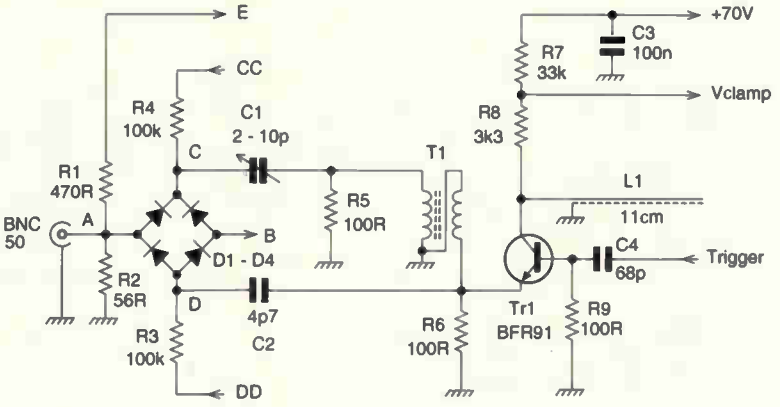
Experiments with sampling

I have long been fascinated by the technology of the sampling oscilloscope, and have felt for some time that it was a field ripe for experimentation.

The really difficult part is a circuit capable of taking exceedingly short samples of the signal. The sampling gate has to open and close within an exceedingly short space of time, to provide the necessary brief sampling 'aperture'. Not only must the gate be capable of very high-speed operation, but also the pulse that operates it must be very narrow. A one-nanosecond wide pulse corresponds to a Nyquist bandwidth of 500MHz.

Sampling oscilloscopes have involved various methods of producing such very narrow pulses. One well known technique uses a 'snap-off diode'. This is a diode which, like most, exhibits stored charge. However, it is so constructed that when reverse biased by a sharp edge, the stored charge is released rapidly, the flow of charge ending very abruptly.

Such diodes are specialised devices. Experiments I conducted some years ago failed to find any commonly available diodes that would work satisfactorily in this mode. So I turned to another technique, which I have touched on in the past, in these pages.[1]



T1, 3T + 3T of 0.008in enameled copper on Siemens 2-hole balun core B62152-A8X30, A_L= 2300

D1 - D4, BAR28. E: to Trigger pick-off amplifier CC - DD: gate G1 back-off voltage.

B: G1 output to sample feedback loop circuitry (see Fig. 3) L1, miniature 50 PTFE insulated coax.

Fig. 1. A few options were tried before this very high speed sampling gate was arrived at.

Avalanche options

Avalanche transistors are of interest again, as fast switches capable of producing short, high-powered pulses to drive semiconductor laser diodes.

For a sampling oscilloscope though, high power is not a consideration. So back in 1996 I experimented to see whether a low power RF transistor could serve to provide the very short pulses needed for a sampling gate.

I found that a BFR91 would operate satisfactorily, giving pulses showing a rise time almost identical with the rise time specification of my Tektronix 475A. This meant that in fact the pulse rise time was something less than 1.4ns, the pulse width being determined by the length of the pulse forming line used.

At that time, I constructed a sampling gate using a quad of BAR28 Schottky diodes and driven by the avalanche pulses. Experiments showed that a signal could indeed be sampled, but that a lot of further work would be necessary to arrive at a viable system.

But other interests intervened, so despite occasional prods from the editor, the prototype was put on the back burner.

A sampler that works

Recently, interest in the prototype was revived, and the necessary ancillary circuitry developed. **Figure 1** shows the sampler as it now stands.

I chose a 50Ω input, since this alleviates many of the problems encountered when aiming at a high impedance input. One of these is 'kick-out', edges from the sampling

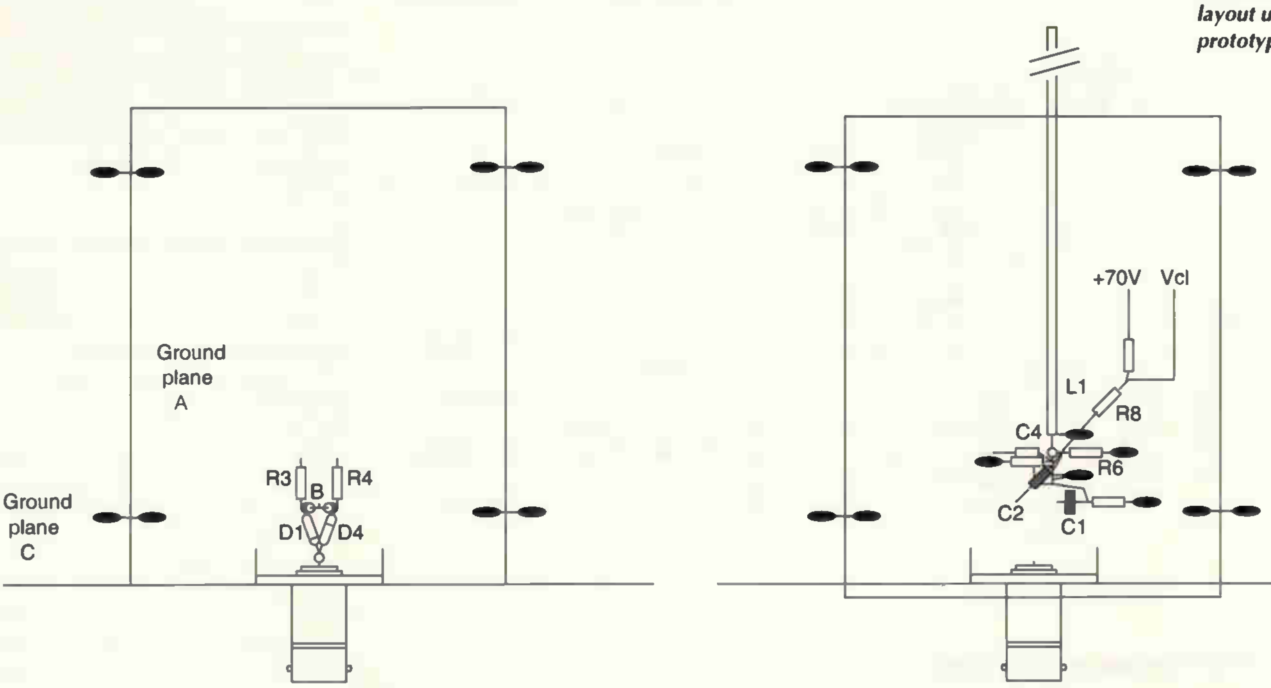


Fig. 2. To achieve best performance, layout of the sampling gate is crucial. These two views of ground plane A show the approximate layout used for prototype.

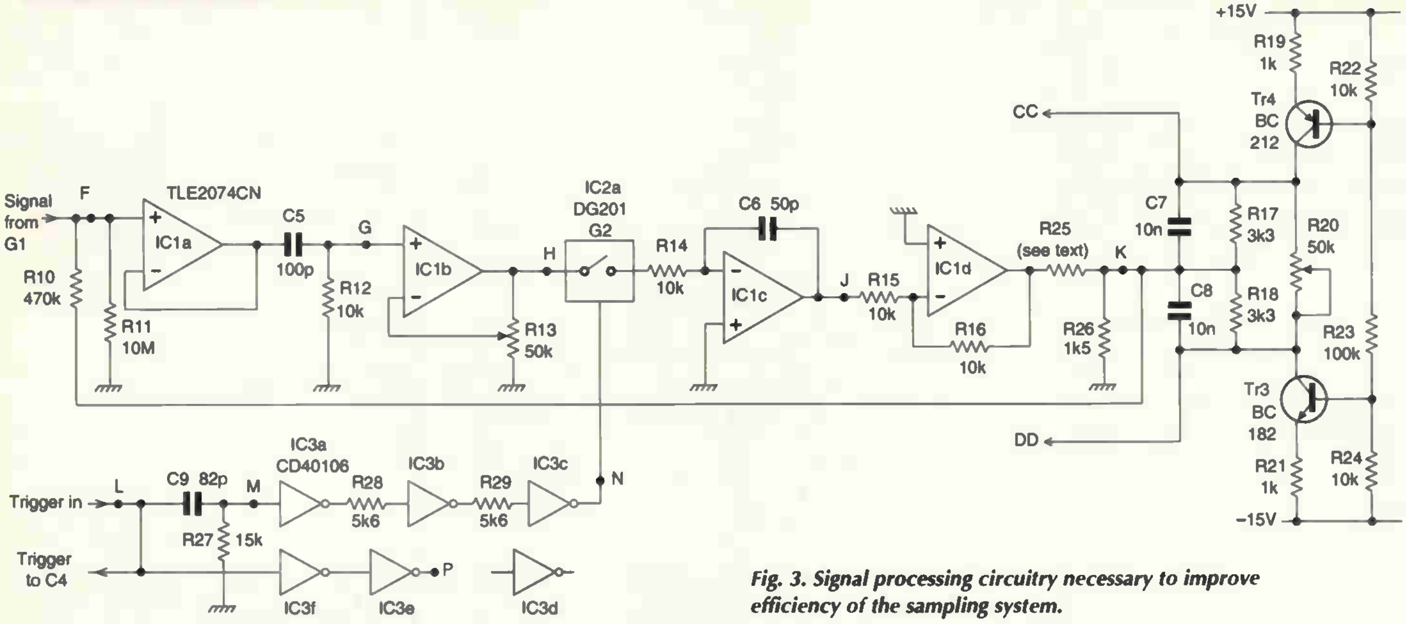


Fig. 3. Signal processing circuitry necessary to improve efficiency of the sampling system.

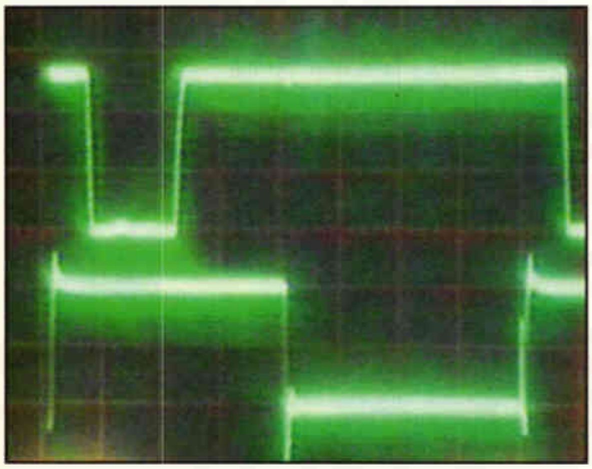


Fig. 4. Lower trace: sample clock at 250kHz, 2V/div, 0V at 1 div from bottom, 0.5μV/div.
Upper trace: negative-going G_2 enable pulse, 2V/div, 0V at centre line.

pulses feeding back into the signal source.

The intention is to drive the input from an active probe with a high input impedance, designs already being to hand.[2] This isolates the circuit under test from kick-out, and incidentally avoids the need to mount the sampling gate in a probe head.

In the absence of a positive-going trigger pulse, the transistor is off, and the 11cm coaxial line at its collector charges up to a defined voltage, V_{clamp}. When a trigger pulse subsequently appears, the transistor goes into avalanche breakdown, becoming, in effect, a low resistance. A positive voltage step therefore appears across R_6, and, via the transformer, a negative step across R_5. This causes diodes $D_{1\text{-}4}$ to conduct, connecting the signal momentarily to point B.

As there are now two 100Ω resistors, a transformer and some conducting diodes effectively in parallel with the line L_1, the collector voltage falls to a voltage somewhat less than half that to which the line was charged.

Collector voltage remains at this value while the voltage step at its input propagates to the open-circuit far end of the line and back again, when it abruptly falls to zero. Thus the width of the pulse is determined by the propagation time of the 22cm round trip on the line.

Assuming the line velocity is 0.66 times the speed of light, the pulse width is theoretically 1.1ns, giving a potential Nyquist bandwidth of 450MHz. Unfortunately, measuring the pulse accurately is way beyond the capabilities of my Tektronix 475A oscilloscope, with its 250MHz bandwidth.

Resistor R_1, at present unused, is earmarked to provide a feed at point E for a fixed ratio prescaler divider, via a wideband amplifier, to provide triggering. Point CC is held positive to point DD, so that in the absence of a sampling pulse, the diodes forming gate G_1 are off.

Implementation considerations

When dealing with very-high frequencies, the mechanical design of a circuit becomes crucial. I built my prototype on a ground plane.

I soldered the square flange of a 50Ω BNC panel-mounting socket to the edge of a 6cm by 8cm piece of SRBP copper clad board (ground-plane A), and strengthened it with a couple of triangular tin-plate gussets, as in **Fig. 2**.

Two of the diodes and resistors $R_{3,4}$ were mounted as shown, on the connector's centre conductor. The other two diodes were mounted pointing upwards, so that point B sits in a hole

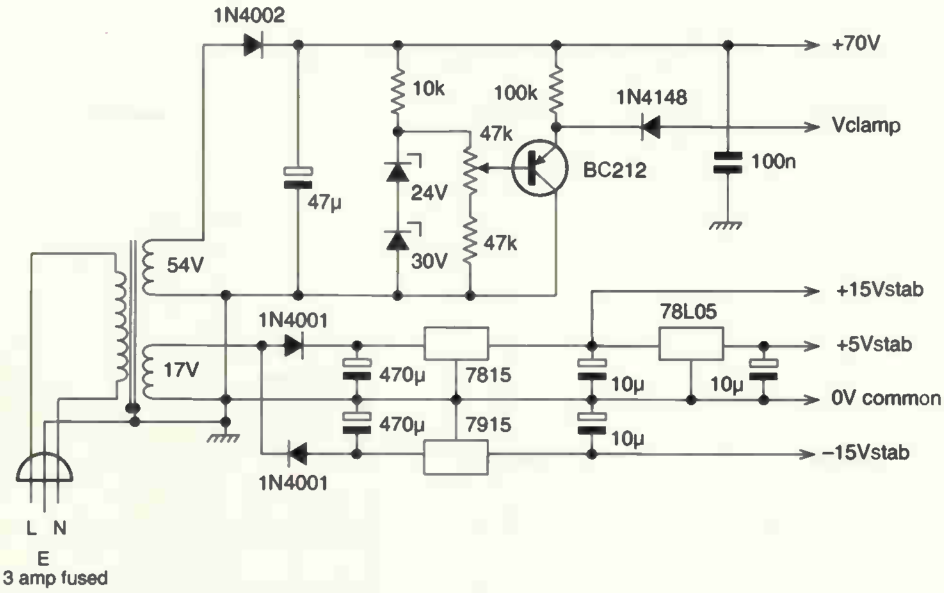


Fig. 5. Suite of stabilised supplies for the system.

in another piece of copper-clad (groundplane B), mounted on metal pillars, above the first.

Remaining components of Fig. 1 were mounted on, or just above the groundplane, as shown separately, for clarity, in the right hand sketch of Fig. 2. The ground plane A was subsequently attached with earth straps to a larger piece of copper-clad, 20cm by 20cm (ground plane C), as indicated.

That completed the critical part of the layout. The beauty of the sampling scope is that once the samples are taken, they can be processed almost at leisure, certainly at lowish frequencies where handling them is no great problem.

Low efficiency

Once the sample is taken, it must somehow be stored ready for display. It can be stored in a capacitor connected to the output of the sampling gate G_1 on Fig. 3, but there is a dilemma here.

If the capacitance is large enough to hold the sampled signal voltage until the next sample, without droop due to the input current of the following stage, it will be so large that a long gate opening will be necessary to charge it. But for a useful system, the gate aperture must be as short as possible. So a very small capacitance is necessary, one which is *not* capable of maintaining the signal until the next sample is taken.

The solution is to use a second gate, G_2, to read the signal sample, the moment it has been taken, into a much larger capacitor. By itself, this simple arrangement is called an open-loop sampling system, and has been used in some early sampling oscilloscopes.

Even with this arrangement though, the short aperture time necessary to achieve a wide bandwidth dictates the use of a sampling pulse so short that the capacitor voltage cannot settle to the full signal voltage. In fact, in a practical design, the 'sampling efficiency' may be only a few percent, a 1V input step being recorded as only a few tens of millivolts.

Lifted by its bootstraps

So a further stage of processing is necessary. Given a sampling efficiency of ε%, say 5%, the sample is fed to an amplifier with a gain of 100/ε, in this case ×20 or 26dB, thereby reproducing the correct full amplitude of the input change.

This replica of what the sampler should ideally have recorded is then stored by G_2 in a large capacitor, in the final stage of the signal-processing circuitry. It is also fed back to G_1, to drag the sampler circuitry up to where it should have got to.

Thus the next sample is ε% of the true change in the input signal voltage, since the preceding one. This closed loop system thus uses sample-loop feedback.

Note that 100/ε gain amplifier must not amplify the absolute input signal voltage, but only its change since the last sample was taken. So before amplification, the signal is differentiated with a 1μs time constant.

The true amplitude difference must

The Achilles heel?

A sampling oscilloscope typically used no input attenuator, being designed to handle a maximum input signal of ±5 or 10V, depending on the model. Such a signal just fills the display in the Y direction. To provide more sensitive Y deflection ranges, it is only necessary to amplify the sample loop output, by ×2, ×5, ×10, or whatever. But there is a limit to this, as explained below.

The performance achieved depends not just on the short aperture time, but also on the positive feedback implemented by the sample-loop feedback circuitry. The loop gain, and hence the feedback has to be just right; not too little, not too much.

In the upper trace of **Fig. A** you can see how a 20kHz waveform sampled at 20kHz displays as a 300Hz waveform; the oscilloscope's horizontal axis was set to 1ms/div. The reconstructed waveform clearly shows the effect of inadequate loop gain in the sample loop feedback circuitry. This is not necessarily a bad thing; indeed in a somewhat less extreme form, it can be useful.

The low loop gain indicates a reduced amount of positive feedback. The latter, as is well known, among other effects, increases noise in a circuit. So a sample-loop feedback setting such as in Figure 12 is useful when examining very small signals, where the signal to noise ratio would otherwise be too small. By greatly increasing the sample density, by displaying only one cycle or less for example, the true rise time can still be measured.

Figure B displays the same waveform (upper trace), and shows the effect of excess loop gain in the sample-loop feedback circuitry. The first sample following a step change overshoots the mark, the second sample over-corrects and so on, the waveform settling to the final value in an underdamped, oscillatory manner.

A further increase in loop gain increases the amplitude of the ring and extends its duration, while with even more loop gain, the sample-loop feedback system becomes unstable. Unlike an underdamped servo system, the frequency of the ringing is easily predicted. It is of course, exactly half the sampling frequency.

The fact remains that to achieve the maximum bandwidth, sample-loop feedback must be adjusted to the optimum, securing exact acquisition of a changed input level on the very first sample. The necessary degree of feedback then inevitably increases the noise level associated with the signal, setting an inherent limit to the smallest signal that can be accurately viewed.

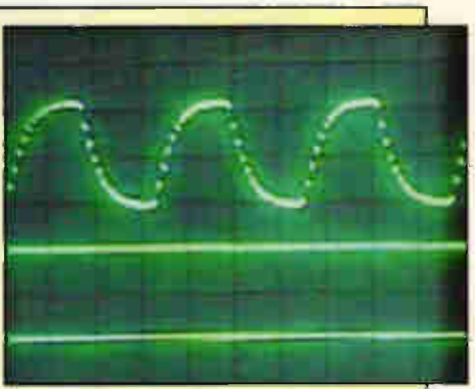


Fig. A. A 20.3kHz waveform sampled at 20.00kHz appears as 300Hz (top trace, 1ms/div trigger source) demonstrating the effect of low sample feedback-loop gain. The 20kHz sampling waveform, lower trace, is not synchronised.

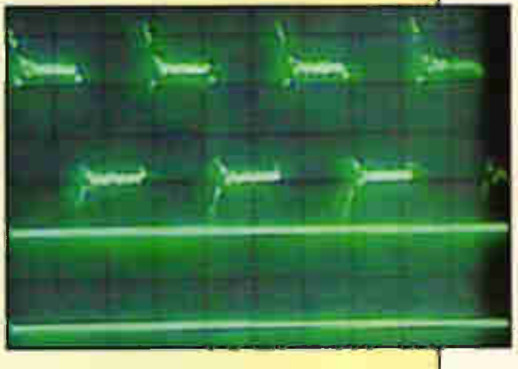


Fig. B. As Fig. A, but showing the effect of excess sample feedback loop gain.

then be added to a store that provides a running memory of the signal voltage. This is achieved by applying it to an integrator. The combination of a differentiator and an integrator makes the loop effectively dc coupled, while permitting corrections for the limited sampling efficiency of gate G_1.

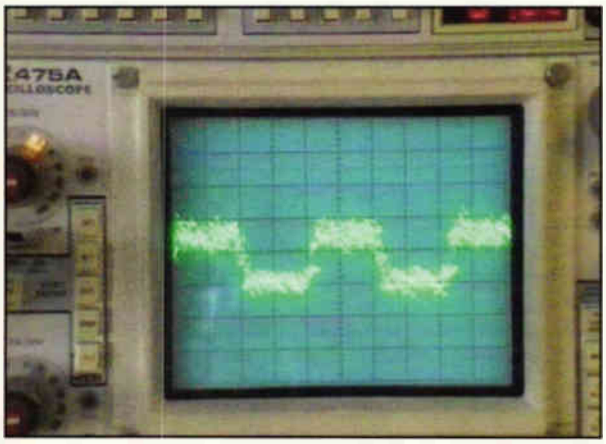


Fig. 6. Jitter on the G_1 sample pulses resulted in a noisy reconstructed waveform.

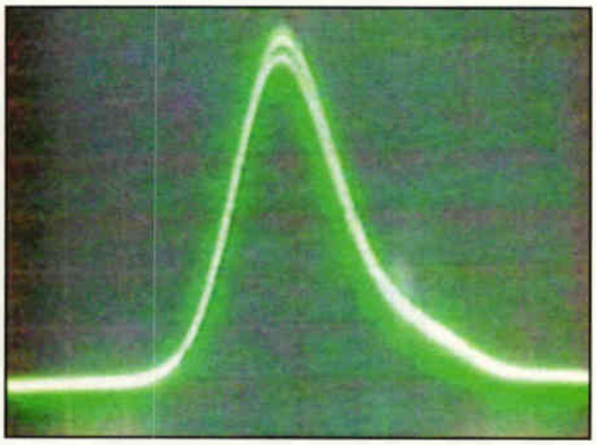


Fig. 7. With a Vclamp of 43V, the pulses from the avalanche transistor are clearly not all the same amplitude. Vertical; >50mV/div. Horizontal; 1ns/div.

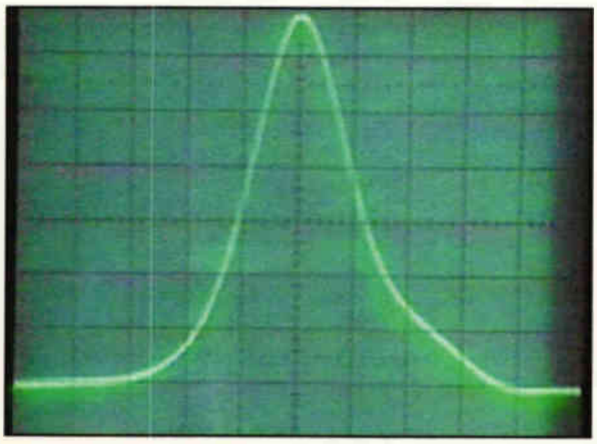


Fig. 8. With V_{clamp} set to 34V, consistent pulses are obtained. Vertical; 50mV/div. Horizontal; 1ns/div.

Figure 3 shows the signal-processing circuitry providing the sample-loop feedback. Comparing this with Fig. 1, you can see that there is in fact no physical capacitor at the output of G_1 for storing the sample in. The capacitance consists of the differential input capacitance at the non-inverting input of IC_{1a}, of some 2.5p, plus the self capacitance of R_{10} and R_{11}, plus circuit strays.

I mounted IC_1, together with $IC_{2,3}$, on a scrap of RS stripboard, mounted on ground plane B. All ICs were socketed.

Components to the right of point K in Fig. 3 were mounted on another scrap of stripboard, mounted at the opposite end of ground plane A from the BNC socket.

How sample-loop feedback works

Assume that the input voltage has been 0V for many samples. The output of G_1 will be zero volts and so will the output of G_2, the other three sections of IC_{2a} being unused.

Output from the signal-processing circuitry at point K is also zero. Now assume that before the next sample, the input changes to +1V. The next sample pulse raises the voltage at point B – the same as point F – to, say +100mV. This voltage appears at point G, and is amplified to +1V at point H.

Very shortly after the sampling pulse, G_2 conducts for a short defined period, chosen in conjunction with the values of R_{14} and C_6, such that the output at point K changes by 1V.

As the integrator is an inverting stage, IC_{1d} is added to make the signal path through all four op amps non-inverting. Non-inverting integrator circuits exist, but these are not so conveniently gated on periodically, as the arrangement shown.

The corrected sample voltage is fed back to the non-inverting input of IC_{1a}, raising the voltage there to +1V. As noted, this is a positive-feedback loop, and could be unstable if not correctly designed.

The importance of timing

I originally intended to run the sample clock at around 100ks/s, so that 1000 points across the screen would correspond to a 100Hz trace refresh rate. However, to permit a clearer display of the gate timings, **Fig. 4** was recorded with the clock rate increased to 250kHz.

The positive-going edge of the sample clock at point L (lower trace) triggers the avalanche transistor, resulting in the G_1 sample pulse, which is not shown.

In addition, the positive-going edge of the sample clock is differentiated by C_9 R_{27} of Fig. 3 and applied to IC_{3a}. The result is a 750ns-wide negative-going G_2 enable pulse at point N, delayed by 250ns on the sample clock and G_1 trigger. This delay allows time for the output of IC_{1b} to settle before G_2 conducts.

During the G_2 enable pulse, the signal-processing circuitry output at point K rises to the full sample value. In turn this starts to raise the voltage at point F from ε% of the input voltage to the full amount.

However, the time constant formed by R_{10} and the capacitance to ground at this point, prevents the voltage rising substantially during the time that G_2 conducts. Thus although the loop feedback is positive, the gain is much less than unity throughout the G_2 enable pulse. In any case, the loop is broken at other times.

Most of the bootstrapping of point F up to the full current sample voltage takes place during the period following the turn off of G_2, during the period between sample pulses.

Total bootstrapping

In addition to centring the output of G_1 on the true current input signal value, ready for the next sample, it is also necessary to centre the positive and negative drive pulses at points D and C on the true current input level. This ensures that, whichever way the signal changes between samples, none of the diodes will be forward biased before the arrival of the next sample pulses.

Constant current generators Tr_4 and Tr_3 produce positive and negative back-off voltages, equally spaced above and below the output of the signal-processing circuit at point K.

Resistor R_{20} provides a means of adjusting the total back-off voltage. The back-off voltages, applied at CC and DD, charge C_1 and C_2 to the appropriate voltage during the between-samples gap, ready to take the next sample.

How the circuit grew

The circuitry was developed stage by stage, starting in early 1996 with the sample gate G_1, as in Fig. 2. The signal-processing block was developed nearly four years later, as a separate exercise. It is driven by a 100kHz square-wave clock.

Originally, I planned to use a TL084 at IC_1, but it proved too slow. In particular, IC_{1b} output took too long to reach its maximum value. This would have required too large a delay on the G2 enable pulse to permit operation at 100ks/s.

So I substituted a TLE2074 instead, which is about three times as fast. With suitably delayed G_2 gating available from IC_3, the signal-processing circuitry was integrated with the G_1 sampling circuitry. At this stage, the back-off voltage generator had not been developed, so R_3 and R_4 were returned to fixed voltages.

As the development progressed, operating the system from a sundry collection of laboratory bench power supplies became ever more inconvenient, so I produced the suite of power supplies shown in **Fig. 5**.

The ±15V, +5V and V_{clamp} supplies were stabilised, the respective raw supplies being derived by half-wave rectification. In view of the low currents drawn at this stage, this was satisfactory, despite some 1.4V pk-pk ripple on the +70V raw supply and 0.7V pk-pk on the others. The supplies were constructed on yet more strip board, mounted, along with the mains transformer, at the rear of ground plane C.

Once the system seemed to be working – albeit far from perfectly – the back-off voltage supplies shown in Fig. 3 were developed and added to the system.

Getting it to work

A 100kHz triggering sample clock, from the TTL output of a Linstead G1000 video oscillator, was applied to point L. It had been intended to trigger the avalanche transistor of G1 from point P of Fig. 3, but the drive from the 40106 proved inadequate, so for now G_1 was triggered from point L.

A signal input to the BNC socket was provided from the variable square-wave output of a second G1000. My work of some four years ago had shown that the G_1 sample output did respond in some measure to a signal input, but with no sample-loop feedback in place, point B showed mainly switching hash.

Now, with the sample-loop feedback in place, an exceedingly ragged, noisy trace resulted, when point K was viewed on an oscilloscope. It was so noisy in fact, that only by varying the input signal frequency slowly, around the same frequency as the sample clock, could it be discerned that there was indeed some response.

A useful improvement resulted from the insertion of an L pad, R_{25} 10kΩ, R_{26} 1.5kΩ, at the output of IC_{1d}, to reduce loop gain. However, the reproduced waveform on the oscilloscope was still very noisy.

I remembered that a noisy, ragged trace was always a problem to some extent when using sampling oscilloscopes. Unfortunately, the amount of noise I was seeing rendered the system unsuitable for any serious use, **Fig. 6**. Here, the sampling frequency was 100kHz.

I set the signal frequency to a slightly higher value, resulting in the 2kHz reconstructed waveform shown. This expedient of offsetting the signal frequency from the sampling frequency – or one of its harmonics – in order to obtain a stable picture, was used because at this stage there is as yet no proper synchronising circuitry.

Initially I suspected 50Hz induced jitter of the avalanche circuit, but 470µF in parallel with the +70V raw supply effected no improvement.

Timing jitter from G_1 still seemed the most likely problem, so I viewed the avalanche pulse generator output directly. The secondary of T_1 was disconnected from R_5, and C_2 from R_6 likewise.

I connected a 50Ω coaxial lead across R_6, the other end of which was connected to the oscilloscope via two 50Ω 10dB pads and a 50Ω through terminator. **Figure 7** shows the sampling pulse recorded.

With V_{clamp} set at 43V, pulses from the avalanche transistor are clearly not all the same amplitude. As viewed directly on the oscilloscope, at least four different amplitudes were apparent.

It was also clear that with the 47kΩ potentiometer in Fig. 5 set to its maximum, V_{clamp} was not reaching the expected value of slightly more than 54V. The answer has to be collector leakage current in Tr_1. After all this transistor is being used way beyond its rating.

Thus the approach of V_{clamp} to a steady value would be exponential. Depending on just where it had got to, triggering might occur marginally sooner or later, giving different height pulses.

Clean pulses – clean traces

The whole point of the V_{clamp} circuitry was to ensure that collector voltage of Tr_1 had stabilised before the appearance of the next trigger pulse. Thus, when the avalanche pulse terminates, Tr_1 collector and line L_1 charge up, aiming at +70V.

Exponential voltage increase is then interrupted when the 1N4148 diode in Fig. 5 turns on, clamping the voltage at the emitter voltage of the BC212. This was clearly not happening, with the results noted. So I set the 47kΩ potentiometer in Fig. 5 to clamp Tr_1 collector voltage at 34V, resulting in clean avalanche pulses, **Fig. 8**.

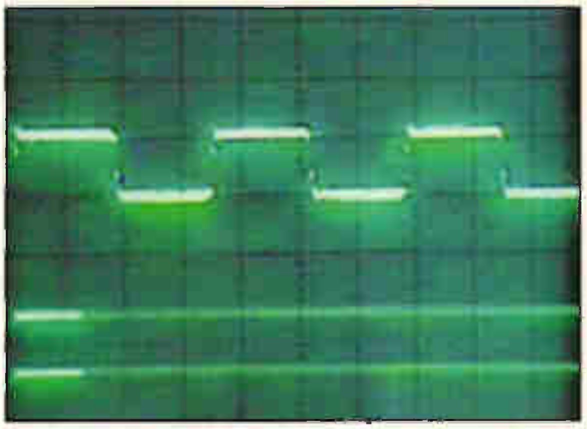


Fig. 9. A 1V 20.35kHz square-wave, sampled at 20.00kHz, is reconstructed as a 350Hz square-wave, upper trace, 1V/div, horizontal 1ms/div. The lower trace, 1V/div, shows the unsynchronised input.

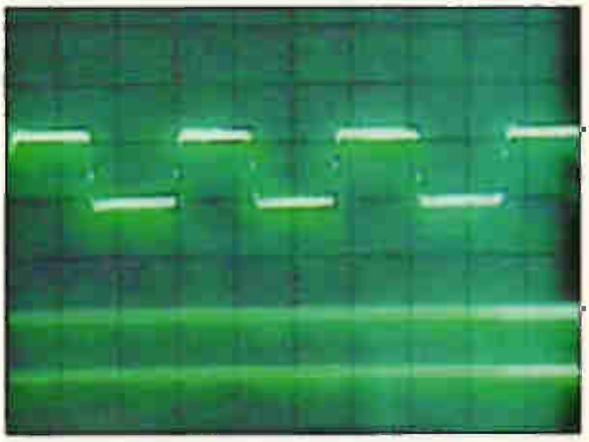


Fig. 10. As Fig. 9, except that the input is a 1V square-wave at about 2MHz.

Allowing for the two 10dB pads, the indicated amplitude is 3.3V across 50Ω in parallel with R_6 and T_1 primary. However, as the calculated pulse width of 1.1ns is less than the specified 1.4ns rise time of my oscilloscope, the actual amplitude is doubtless greater than this, by a margin that's anybody's guess.

Clean pulses at last

With clean pulses now available, the reconstructed waveform was greatly improved. **Figure 9** shows 1V positive-going 50/50 mark/space pulses at 20.30kHz. The upper trace, now sampling at 20.00kHz rather than 100kHz as previously, shows the reconstructed trace at 1V/division, 0V ground at the centre line, 1ms/div. This results in an apparent frequency of 300Hz, from which the display is triggered.

The lower trace, also at 1V/division, 0V ground at 1 division up from bottom, is the input signal, not synchronised. You can see that, thanks to the sample-loop feedback, the system acquires the amplitude of the signal, following a transition, almost completely on the first sample following. In fact, the apparent shortfall is due to the sample actually occurring during the rise or fall-time of the pulse.

Figure 10 shows the same thing, except that the input frequency is now about 2MHz. The actual frequency is 350Hz higher than the nearest harmonic to 2MHz of the 20kHz sampling waveform. There is clearly very little difference.

The beauty of a sampling oscilloscope is that, provided the width of the sampling pulse – the aperture time – is short compared with the half period of the waveform being viewed, the frequency of the latter is immaterial.

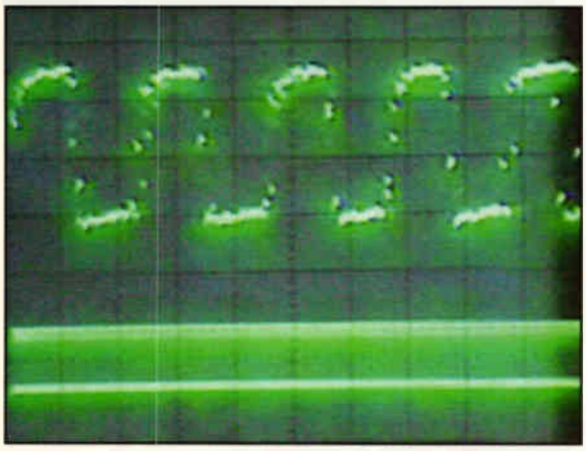


Fig. 11. Trace 2 showing an unsynchronised 1V 10MHz input (1V/div.), trace 1 (upper) showing the reconstructed waveform at 2V/div., monitored at IC_{1d} output, i.e. ahead of the $R_{25,26}$ pad.

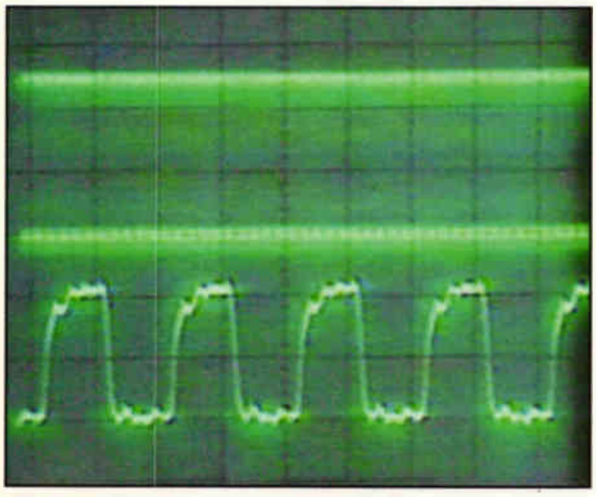


Fig. 12. As Fig. 11, except triggered from Trace 2, now at 0.5V/div., showing the 10MHz input waveform, warts and all.

Recording the reconstructed waveform in Fig. 10 posed a problem. The G1000 oscillators are basically Wien-bridge circuits. Like all *CR* oscillators, their phase noise is worse than that of an *LC* oscillator – and of course much worse that of a crystal oscillator. But the sampler G_1 is sampling only every hundredth pulse of the signal, resulting in a rather wobbly trace. This prevented the use of my home-made oscilloscope camera, since it requires a 20 second exposure.

So a camcorder was set up in front of the oscilloscope, and a recording several seconds long made. This was then played into the computer using a Motion Picture 1040-24 Colour interface card.[3]

After a few attempts, a frame showing a stable picture was captured, as reproduced here. All the other screen shots were made in the same way.

What bandwidth?

In Fig. 10, samples can be seen to fall at different points up or down the edges of the input waveform.

With results so encouraging, I was able to reduce the 18dB attenuation in the sample-loop feedback to further enhance performance. This attenuation was added in the interests of stability, before the G_1 pulse jitter was cured. To this end, R_{25} was reduced from 10kΩ to 4.7kΩ. With this modification, the input signal was set to about 10MHz, and sampled as before at 20kHz.

Figure 11 shows the reconstructed trace, channel 1, monitored at the output of IC_{1d} at 2V per division, and the input 10MHz waveform, channel 2, lower trace unsynchronised, at 1V per division. This may be compared with the input signal itself, **Figure 12** lower trace at 0.5V per division, triggered from the input signal. Apart from a slight slope on the 0V level of the signal, at present unexplained, the sampled signal is a remarkably faithful reproduction of the original, even down to the ripple on the +1V level.

While I do not at present have the instrumentation to determine the sampling system's bandwidth or rise time, the calculated sampling pulse width of 1.1ns predicts a potential Nyquist bandwidth of 450MHz. And it is clear from the correspondence between Figures 11 and 12 that in practice, the bandwidth must extend well into the VHF region, possibly the UHF.

Further work

The results achieved so far are sufficiently encouraging to warrant further work.

In addition to further optimising the sample-loop feedback provided by the signal-processing circuitry, the major missing part of the jigsaw at present is the triggering department.

Briefly, this will use a fast and a slow ramp, as in any sampling oscilloscope. The latter will feed out to the external X input of an oscilloscope, while the former is triggered by a divided down version of the input signal.

A comparator detects the instant at which the fast ramp voltage passes that of the slow ramp, initiating a sampling pulse at that instant. As the slow ramp voltage rises, sample pulses are produced with a steadily increasing delay relative to the corresponding trigger pulses, building up a reconstructed picture of the input signal, similar to those shown.

By attenuating the version of the slow ramp fed to the comparator, the additional delay per sample is reduced. Ultimately, if the slow ramp becomes effectively horizontal, all samples occur at the same point on the waveform. This corresponds to an infinite sample density, so an impressive figure of so many (effective) giga samples per second is easily achieved! Readers wishing a more detailed description of the sampling scope will find it in chapter 6 of reference 4.

Some work on this triggering/sample density department was done along with the earlier sampling circuit development, so I hope it will not be too long before a complete design for a wideband-sampling adapter is available. This would extend the capability of any modest oscilloscope to hundreds of megahertz – at least for repetitive signals. ■

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- When a quick indication of the input signal is required, a simple click on the auto setup button will immediately give a good overview of the signal. The auto setup function ensures a proper setup of the time base, the trigger levels and the input sensitivities.

- The sophisticated cursor read outs have 21 possible read outs. Besides the usual read outs, like voltage and time, also quantities like rise time and frequency are displayed.

- Measured signals and instrument settings can be saved on disk.This enables the creation of a library of measured signals. Text balloons can be added to a signal, for special comments. The (colour) print outs can be supplied with three common text lines (e.g. company info) en three lines with measurement specific information.

- The HS801 has an 8 bit resolution and a maximum sampling speed of 100 MHz. The input range is 0.1 volt full scale to 80 volt full scale. The record length is 32K/64K samples. The AWG has a 10 bit resolution and a sample speed of 25 MHz.The HS801 is connected to the parallel printer port of a computer.

- The minimum system requirement is a PC with a 486 processor and 8 Mbyte RAM available. The software runs in Windows 3.xx / 95 / 98 or Windows NT and DOS 3.3 or higher.

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CIRCLE NO. 118 ON REPLY CARD

Illustration Anna-Louise Barclay

Hands-on Internet

Photodiode sensing

Using the Internet, Cyril Bateman has been examining the benefits of transimpedance amplifiers over transconductance amplifiers when measuring light using a photodiode.

As I mentioned in my article in the January issue, a design I was working on late last year involved converting a voltage input into a current output.

While searching the Internet for operational transconductance amplifiers suitable for carrying out the conversion, I found many claims of much improved bandwidth from using current instead of voltage to drive rectifying diodes, optocouplers and inductive components.

The inverse situation, where an amplifier that accepts an input current to produce an output voltage, is called a 'transimpedance amplifier'. It seemed reasonable to expect similarly improved results using transimpedance amplifiers to monitor the output from these components.

Search results

The Global Semiconductor Datasheet site[1] found only 27 references to 'transimpedance', and

Where to look...

1.	Global Semiconductor Datasheet library	http://www.semi.com.tw
2	The Engineer's Room	http://www.anadigics.com/engineers/TIA.html
3	Optoelectronics Designers Catalogue	Hewlett Packard Components.
4	Photodiode Monitoring with Op-Amps. AB075.PDF	http://www.burr-brown.com
5	Keeping the Signal Clean in Photosensing Instrumentation	http://www.sensorsmag.com/articles/0697/photo_s/main.shtml
6	Kodak Wratten Filters	Pub. Kodak Ltd.
7	Maxim Integrated Products	http://www.maxim-ic.com
8	Analog Devices Inc.	http://www.analog.com
9	Philips Semiconductors	http://www-us.semiconductors.philips.com

none for the term 'transresistance'. By comparison, searching Alta Vista and Google I found more than 1000 hits.

Many of these hits were from smaller integrated circuit makers not listed on Global. Others were from equipment manufacturers and Universities. I found lower frequency transimpedance amplifier circuits could be built using conventional FET input op-amps.

Using nothing more elaborate than a 10kΩ resistor, transconductance techniques allow you to convert a current of 1μA to voltage and monitor it on a DVM. Currents smaller that 1μA are rather more difficult to measure in this way though. Transimpedance amplifier techniques on the other hand can be used to monitor extremely small currents.

Measuring light

The most common application of transimpedance is measuring light levels by monitoring the output from a suitable photodiode. Temperature is the most frequently measured physical parameter, and monitoring light levels is the second most common requirement.

Light energy impinging on a photodiode can be quantified by measuring either the voltage or current output of the diode.[3] So if a photodiode produces a voltage across its terminals that varies with light level, why is its current output normally measured using a transimpedance amplifier?

When connected as a photo-voltaic source, a photodiode's sensitivity varies with its terminal voltage. Voltage output by an op-amp connected to the photodiode represents a logarithmic response to the light level incident on the diode – even though no current is taken from the photodiode[3]. **Fig. 1a)**

Light on a linear scale

While many applications need a logarithmic response to light level, many are better served by a response that increases linearly with incident light. A linear response results when the diode's output current is directed into either a zero impedance or an extremely low input impedance relative to the diode's impedance. This current can be converted to a voltage output, using a transimpedance amplifier.

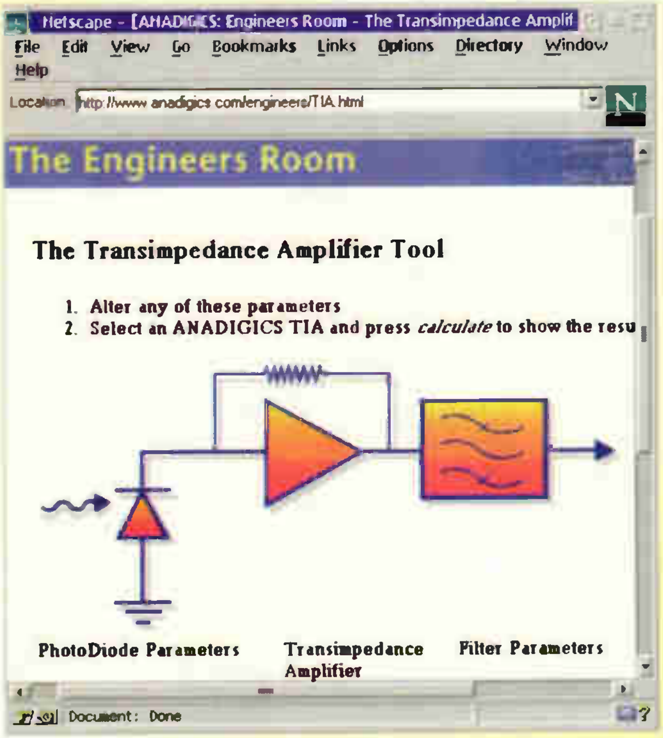
A transimpedance amplifier is easily arranged by connecting the diode into an op-amp's virtual-ground input circuit. Using an amplifier with a very high open-loop gain ensures a minimal voltage can develop across the photodiode.

Current from the diode passes through the amplifier's feedback resistor to the amplifier. The amplifier output is a voltage equal to the diode current multiplied by the feedback resistance value. Ideally, no diode current is taken by the amplifier's input.[4]

While the value of the feedback resistor used might be very large, division by the amplifier's open-loop gain ensures that the diode feeds into a low input impedance

Free photodidode design tool

Anadigics' web site[2] provides a simple design tool that allows 'what-if' calculations to be performed on-line. You simply enter your choice of photodiode, transconductance amplifier and filter characteristics and click on the 'calculate' button. The page displays the bandwidth, sensitivity and optical overload capacity in dBm, of your design, **Fig. A.**



Netscape - [ANADIGICS: Engineers Room - The Transimpedance Amplif

File Edit View Go Bookmarks Links Options Directory Window Help
Location: http://www.anadigics.com/engineers/TIA.html

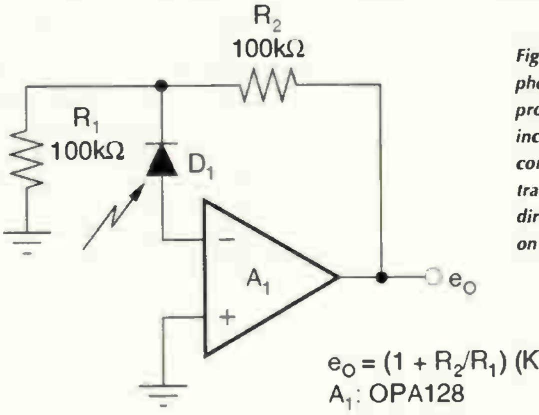
The Engineers Room

The Transimpedance Amplifier Tool

1. Alter any of these parameters
2. Select an ANADIGICS TIA and press *calculate* to show the resu

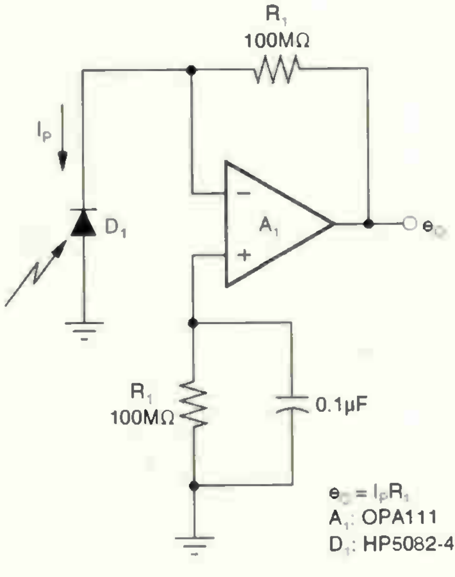
PhotoDiode Parameters
Transimpedance Amplifier
Filter Parameters
Document: Done

Fig. A. Although restricted to Anadigics TIA part numbers, this Javascript program allows quick and easy 'what if' design calculations.



$e_O = (1 + R_2/R_1)(KT/q)\ln(1 + I_P/I_S)$
A_1: OPA128
D_1: HP5082-4204

Fig. 1. With its photodiode connected in photo-voltaic mode, circuit a) outputs a voltage proportional to the logarithm of the light intensity incident on the diode. In b), the diode is now connected in the photo-conductive mode so the transimpedance amplifier outputs a voltage directly proportional to the light intensity incident on the diode.



$e_O = I_P R_1$
A_1: OPA111
D_1: HP5082-4204

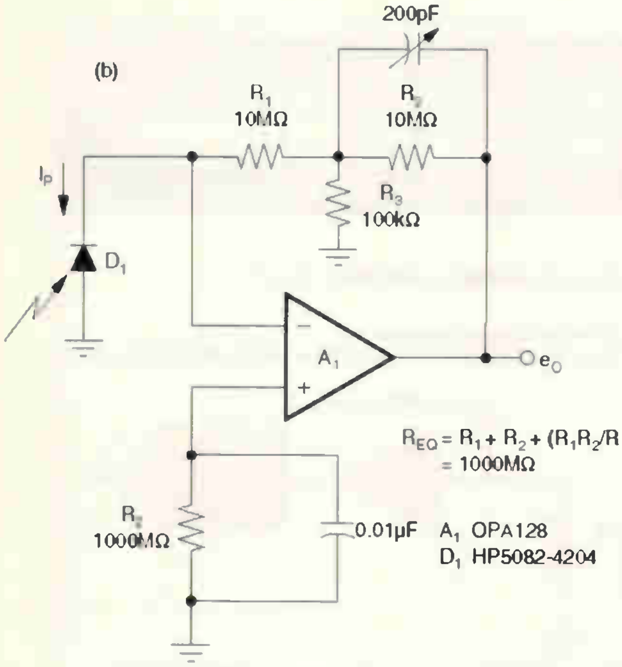


Fig. 2. This transimpedance amplifier using a 'T' feedback network of low-value resistors produces similar gain to one using a single 1000MΩ feedback resistor. The 200 pF trimmer can be adjusted to eliminate high-frequency gain peaking.

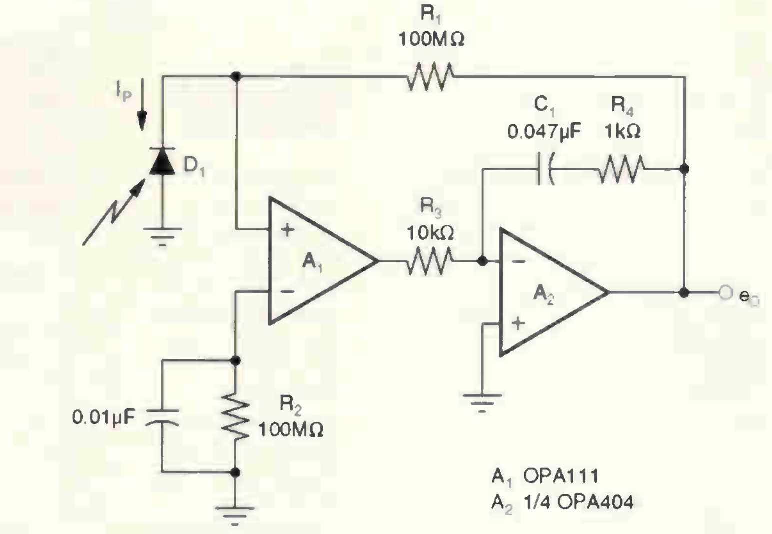


Fig. 3. This composite transimpedance amplifier reduces high frequency noise. It provides a low frequency transimpedance of 100MΩ reducing effectively to 10MΩ at higher frequency.

compared to its own output impedance.

Voltage across the diode should be as near zero as possible. Using a high input impedance, low bias current, FET amplifier, this circuit provides a linear response, **Fig. 1b**).

Minimising strays and noise

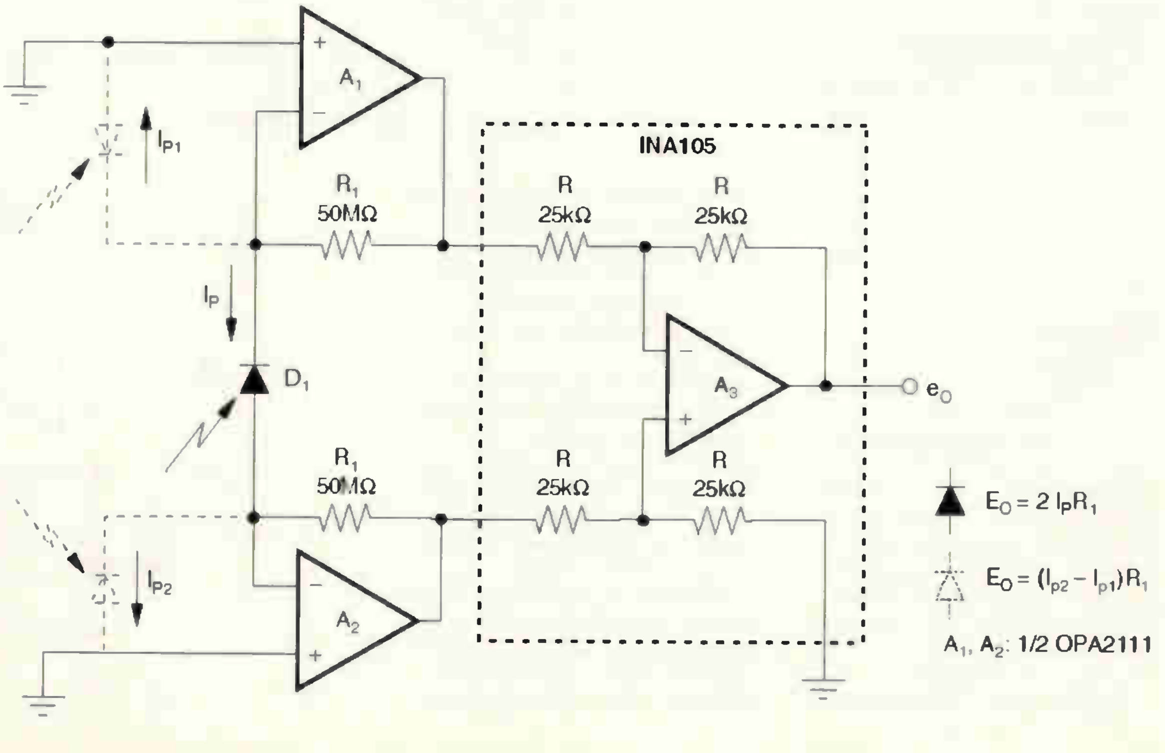
Such high value feedback resistors pose two difficulties, increased circuit noise levels and reduced bandwidth. The extremely high resistor values render the circuit susceptible to variations in stray capacitance and surface leakage. Shielding against noise pickup and printed board guard tracks to protect against surface leakage currents are essential precautions.

The normal op-amp pinning, with the negative supply to pin 4, offset trim to pin 1 and inputs to pins 2 and 3, does not help.^{5} Using a TO-99 package and tying its input pins to PTFE stand offs can dramatically reduce the affects of surface leakage currents.

Using plastic DIP or SOIC packages with efficient guard tracks is much easier when the op-amp has a non-standard pin-out. One such op-amp, the Burr-Brown OPA129, features an ultra-low bias current of 100fA (femtoamps). Having no connections to its pins 1 and 4, facilitates providing the essential printed circuit board guard tracks. The negative power supply for this device is wired to pin 5.

A resistor 'T' feedback network circuit can be used to effectively replicate the very high value feedback resistor, but using more modest resistance values. A small

Fig. 4. Basic instrumentation amplifier circuit offering two modes. As shown with D1 it provides gain similar to that of a single amplifier circuit using 100MΩ but using only 50MΩ resistors. Alternatively, deleting D1 and using the two diodes shown in dotted outline, it provides the difference between two light levels.



variable capacitor can then be added across part of this 'T' to control gain and noise peaking. Such peaking also occurs using a single high value feedback resistor. Note that this 'T' technique cannot be applied to the bias resistor. **Fig. 2**.

Noise reduction

When amplifying very low input currents, it may be necessary to reduce overall noise sensitivity without reducing the usable bandwidth. This may be accomplished using phase compensation to reduce unwanted high frequency gain.

Out-of-band noise can be reduced by inserting a phase compensating op-amp within the main amplifier's closed loop, making a composite amplifier.[4] To allow this composite amplifier to maintain a single phase inversion, it is necessary to reverse the inverting and non-inverting input connections at the main amplifier.

At DC and low frequencies, the feedback of the added amplifier is blocked by C_1 and the overall open loop gain is the product of both amplifier gains. At higher frequencies, the gain of A_2 becomes less than unity, resulting in less gain overall than from the main amplifier A_1, reducing high frequency noise. **Fig. 3**.

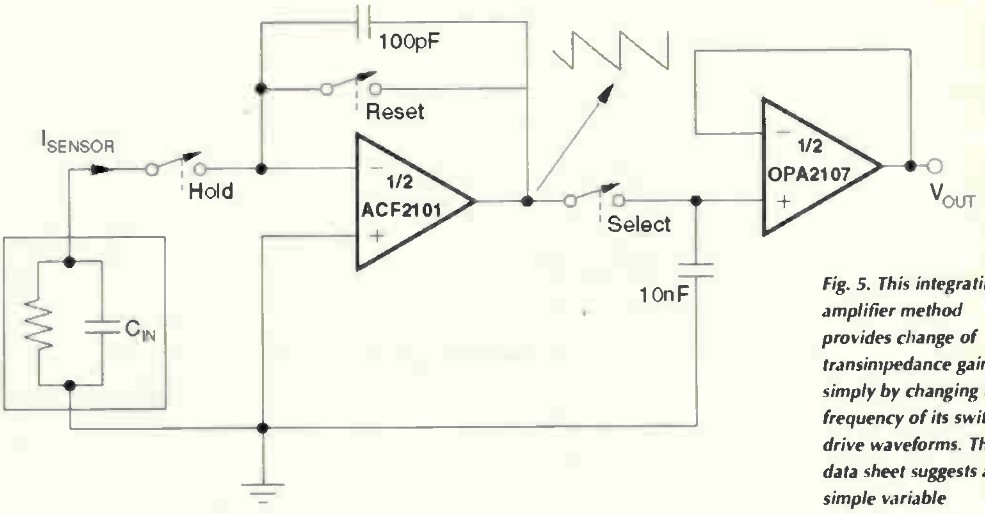


Fig. 5. This integrating amplifier method provides change of transimpedance gain simply by changing the frequency of its switch drive waveforms. The data sheet suggests a simple variable frequency generator, based on three 555 timer circuits, which can be used with the ACF2101, to make a programmable transimpedance

An instrument-amplifier approach

A variation of the instrument amplifier technique can allow half the usual resistance values to be used, yet produces the same overall gain.[4]

The normal instrument amplifier inputs are grounded, the photodiode replacing the conventional gain setting resistor. Using high-value feedback resistors, both input amplifiers are transimpedance amplifiers. Common-mode signals and the input bias current error voltages for A_1 A_2, are rejected by A_3. **Fig. 4**.

Perhaps you need to measure the difference between two currents or two light intensities. Remove D_1 and insert the two comparison currents or photodiodes as shown in dotted outline. The output voltage represents the difference between these two currents.

An integrating amplifier method

With extremely small diode currents, the noise output from a FET op-amp transimpedance amplifier might be unacceptable. It

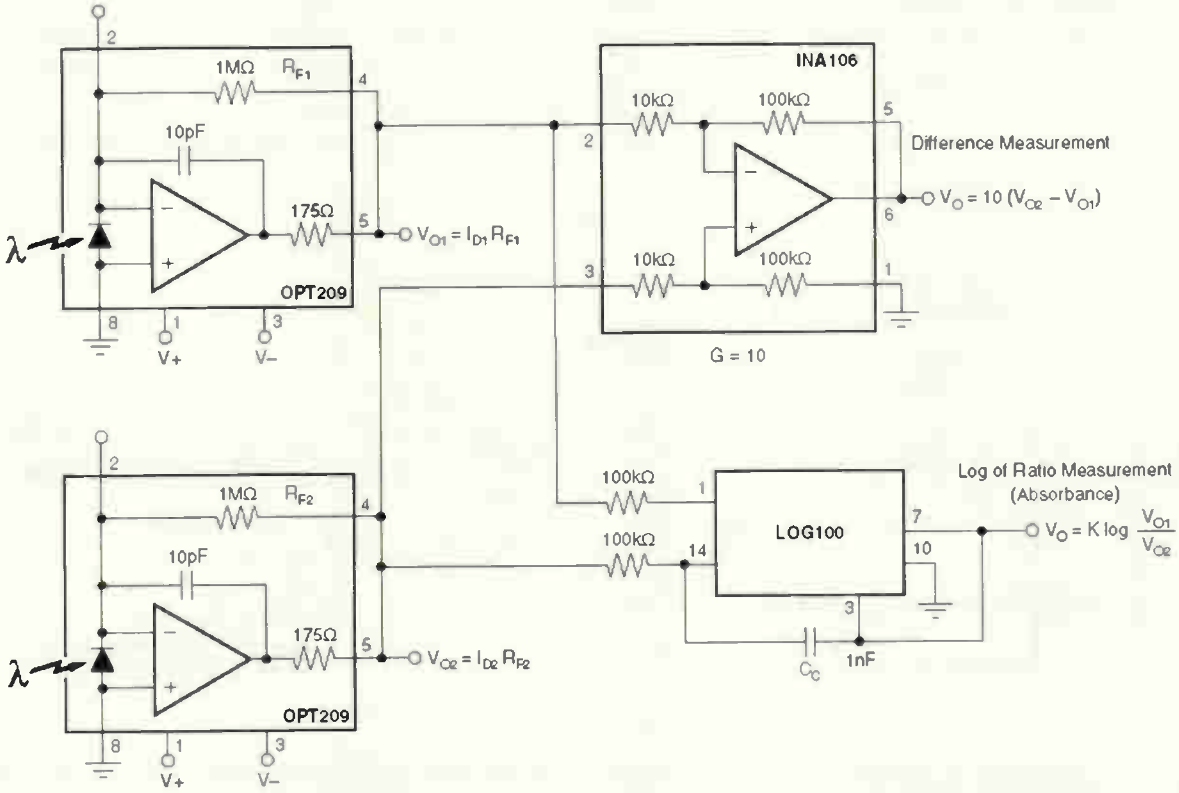

2
1MΩ
R_{F1}
4
10pF
175Ω
5
V_{O1} = I_{D1} R_{F1}
OPT209
8
1
3
V+
V–
INA106
10kΩ
100kΩ
2
3
5
6
1
Difference Measurement
V_O = 10 (V_{O2} – V_{O1})
G = 10
1MΩ
R_{F2}
V_{O2} = I_{D2} R_{F2}
LOG100
14
7
10
C_C
1nF
Log of Ratio Measurement
(Absorbance)
V_O = K log V_{O1}/V_{O2}

Fig. 6. Using two pre-packaged photodiodes and amplifiers to measure the absorbance 'D' of a material.

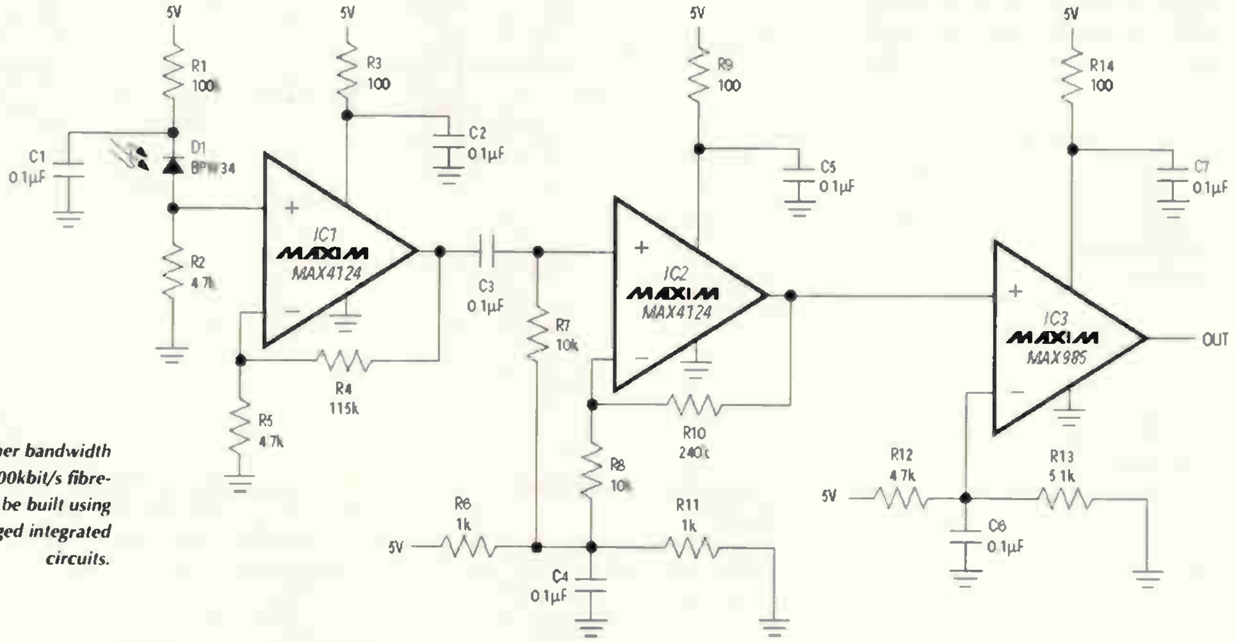


Fig. 7. For higher bandwidth applications, this 800kbit/s fibre-optic receiver can be built using standard packaged integrated circuits.

may be possible to improve noise performance using an integrating amplifier technique.

Burr-Brown's ACF2101 is a dual low-noise switched integrator with 120dB dynamic range, 100fA bias current and 10µV rms noise level.

By varying the device's switching frequency, it can be made to simulate a transimpedance amplifier with a feedback resistor varying from 1MΩ to 1GΩ. Gain of the device can be varied, resulting in a programmable *I* to *V* converter.

No high value resistor is used, so the AFC2101 maintains excellent signal-to-noise ratio and bandwidth – which extends to half switching frequency. **Fig. 5**.

Integrated photodiodes

Many semiconductor makers provide pre-packaged solutions of a photodiode, transimpedance amplifier and feedback resistor.

Some packages allow variation of this feedback resistor value by inserting additional resistance in series with the internal resistor. Others have an extra pin to use an external resistor, the internal resistor being disconnected.

In the datasheet for the OPT209 'photodiode with on-chip amplifier', Burr-Brown shows how to use two OPT209 packages with a LOG100 logarithmic amplifier to measure log ratio or absorbance. This is commonly referred to as the 'D' or optical density of a material.[6] **Fig. 6**.

The above circuits can provide excellent sensitivity and linearity but at relatively low repetition frequencies. For example, the OPT209 using its internal feedback resistor, has a 16kHz

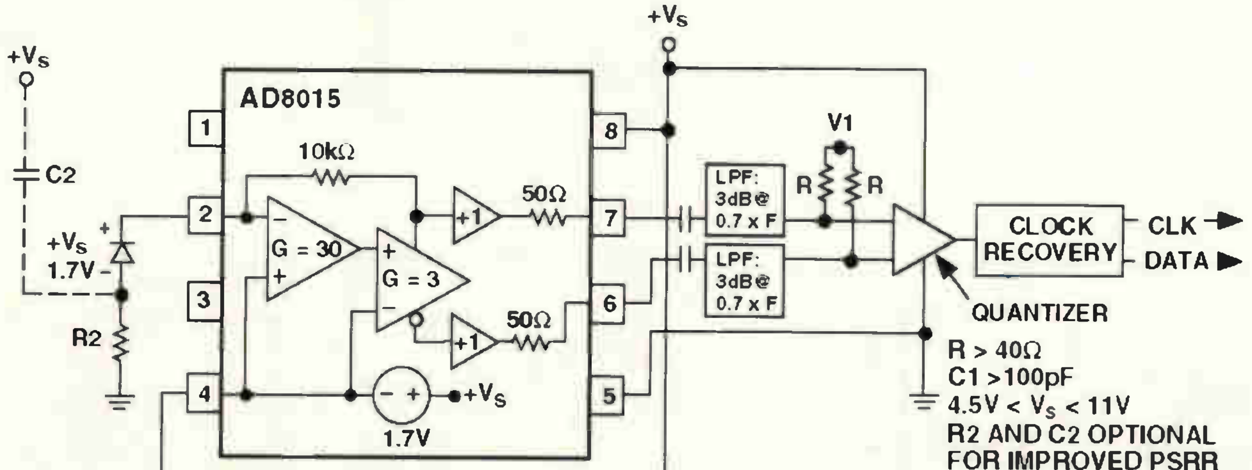


Fig. 8. One implementation of a high-speed fibre-optic receiver. The biased photodiode can be referred either to ground as shown, or to the positive supply rail. Ground referencing increases the diode's bias voltage, reducing its self capacitance and maximising bandwidth.

bandwidth. Fibre optic cable data transfer may need much higher repetition frequencies.

Increasing bandwidth

One way to increase bandwidth is to use a small transimpedance gain, then add additional voltage gain stages as needed, to restore sensitivity. Inevitably, this approach increases system noise.

The very highest data rates demand custom chip design and miniature surface mount packages. For intermediate data rates though, standard components can be used.

One such method is described in Maxim's application note A3015.PDF.[7] This shows how to build an 800kbit/s fibre-optic receiver using a photodiode with two fast op-amps and a comparator.

This circuit uses a 4.7kΩ resistor in series with a biased diode to convert the diode current to voltage. This voltage is applied in turn to two stages each providing a gain of 25. Combined with the 4.7kΩ resistor, this represents a transimpedance of almost 3MΩ, with a bandwidth able to support 800kbit/s.

Using a single transimpedance amplifier with a 3MΩ feedback resistor could provide the same sensitivity but at lower frequencies, **Fig. 7**.

In 1994, Analog Devices[8] announced its first Silicon transimpedance integrated circuit able to meet the bandwidth, noise and sensitivity specifications of a 155Mbit/s data link. Supplied as either an 8-pin SOIC package or as a die, the AD8015 is usable over the extended industrial temperature range of –40 to +85°C.

For these data rates it is essential that the photodiode's self capacitance and the circuit's input capacitance are minimised. The AD8015 input pin in the SOIC package has 0.4pF total stray capacitance and the typical photodiode, when used with a 1.7 V bias, contributes 0.3pF, **Fig. 8**.

Application note AN1435 from Philips Semiconductors[9], describes a family of wideband low-noise transimpedance amplifiers with differential outputs. These are supplied with a choice of feedback resistor, allowing a trade off between bandwidth and input impedance.

The companion application note AN1431[9] describes the design of a SONET receiver built using two integrated circuits – an SA5223 and an NE5224. This note includes the schematic circuit and printed board layout used, together with performance measurements of the prototype. ■

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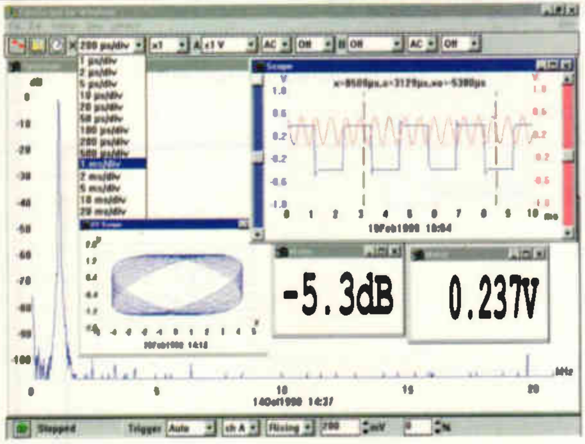

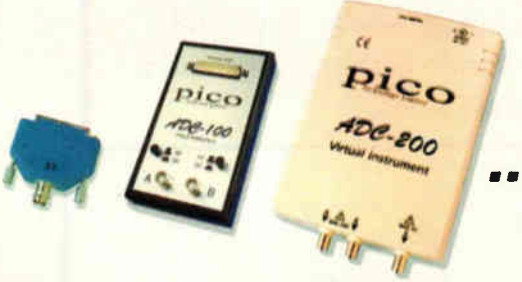

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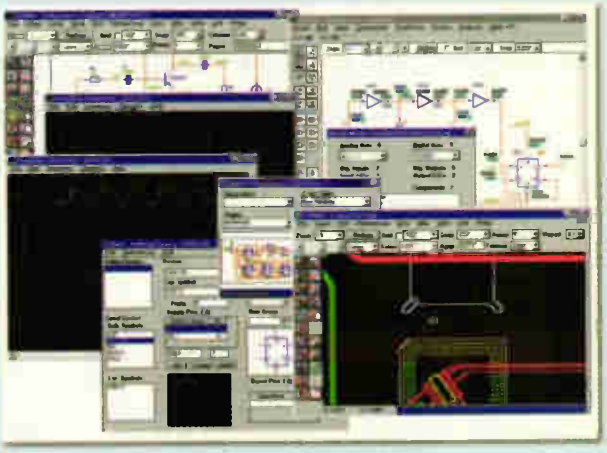

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JLH

a lifetime in electronics

John Linsley Hood looks back at his evolution from an eight-year old wondering how a crystal set worked, to a renowned analogue electronic design guru. This first article looks at his early wireless encounters.

Often a small, apparently insignificant, act can alter the whole direction of future events. I was an eight year old schoolboy when I first experienced such a turning point, when my grandfather, a retired chemistry lecturer, unearthed an old crystal radio from his loft. It was complete with a pair of Ericsson headphones and a crumpled silver cat's whisker – but no crystal and no aerial, **Fig. 1**.

The lack of an aerial and crystal was soon satisfied. My local Woolworths sold plastic-insulated stranded copper 'aerial wire' at sixpence for fifty feet and my grandfather retrieved a small lump of galena (lead sulphide) from his collection of geological specimens.

My newly acquired treasure consisted of a rectangular mahogany box with an engraved ebonite lid. On it were mounted four massive brass terminals, together with two large pivoted coils. Inside the box, there was an interleaved parallel flat-plate tuning capacitor, with a large black knob.

Having got it home, I connected it to the 'earth' – a black painted drain pipe outside my bedroom window – the headphones, and an aerial strung out across the garden. I was rewarded with complete silence.

Confronting this situation, I reasoned that the receiver must have worked when it was new. I set to work renovating it with shellac varnish, oil and metal polish. Having improved my aerial and earth, I then adjusted my expectations downwards and I tried again. As I probed the surface of the crystal, my persistence was occasionally rewarded with faint elfin voices and fairy music.

Although he professed to no real knowledge of this topic, my grandfather said that he remembered that everything depended on the crystal, and that to get the best results it really needed to be mounted in a metal cup containing a blob of some low melting point solder such as Wood's metal or Rose's alloy. He thought he might have just such a thing in his box of odds and ends.

My grandfather also recalled from the early days of 'wireless' that some experimentally-minded enthusiasts had obtained good results when they had used all sorts of strange materials as the 'crystal' – but they all must, of course, conduct electricity.

Thus encouraged and re-equipped, I began a programme of research into 'crystals' and 'cat's whiskers'. To my great satisfaction, I found, that a spiral of plain copper wire in contact with a small piece of coke from the domestic coal shed was by far the best. The wire had to have a clean point – ideally one that had been freshly hammered flat, which made the copper wire stiffer and more inclined to stay in place.

A second influence on my destiny

The final, event that shaped my destiny in the field of electronics happened during the following summer. My brother and I were taken on holiday to stay with my uncle. He had a farm at a small village called Pity Me, just north of Durham.

The farm was mostly arable, and was a fascinating place to us 'townies'. It was of about a hundred and fifty acres in extent, and lay astride the main LNER east-coast railway line from London to Newcastle.

Down in a slight valley, a few hundred yards below the farm house was a brown and yellow painted signal box serving the main line, as well as a local line from Kimblesworth colliery.

My brother, who was two years older than me and a lot more enterprising, soon visited the signal box. He became acquainted with two of the signalmen, one of whom, a Mr Bare – 'Auld Nekkid', as my brother nick-named him in the local dialect – turned out to be of great interest to me. He was the first person I had met who was knowledgeable about 'wireless', and might have some thoughts to add on my crystal radio experiments.

Mr Bare was a radio enthusiast. He showed me how to read a circuit diagram. To my great delight he also gave me a circuit diagram, together with the components I would need to build a one-valve short-wave radio. He said it would work fine with my headphones,

The once common crystal set. Available in the early 1920s, this version needed a long antenna.

and that I should be able to pick up signals from America.

My father found me a small soldering iron that could be heated to working temperature on a gas ring. Astonishingly. Marks and Spencers proved to be the cheapest source of a two volt 'accumulator' and a one hundred and twenty volt 'HT' battery.

I assembled the components of the radio as shown in Mr Bare's circuit diagram. **Fig. 2**. Luckily. as it turned out. Mr Bare's radio design didn't work. though by fiddling with it one could get various clicks and hums.

In the pursuit of knowledge. I explored my local library which was in those expansive days full of books on radio. I soon found the problem – the 'reaction' coil. L_3 in Mr Bare's diagram. was connected to the wrong end of the RF choke in the anode circuit of V_1.

When the connection was corrected. the radio did indeed work fine. Its ability to pick up any signals at all though depended on the correct setting of the reaction coil. which was very delicate. It changed with the tuning and would make the radio whistle loudly if set too high.

However. my studies in the library had expanded my knowledge *and* my ambitions. I discovered that Reading. the town where I lived at the time. had a number of 'junk' shops where one could buy all sorts of outdated radio components cheaply. By the time of my next visit to my uncle's farm I was the proud owner of a home-made three-valve radio. shown in **Fig. 3**.

This new radio had a push-pull output. and would drive a small. baffle-mounted loudspeaker to a good volume. My mother. who liked novelties. was fascinated to be able to listen with me to a news broadcast from RCA in New York. My father merely asked if I could stop it whistling.

To my regret. when I told Mr Bare of my adventures. in the hope of a comprehensive technical discussion on radio. he merely said. 'there's nothing I can tell you. lad. you know a lot more than I do'.

Apart from the need. periodically. to replace the HT batteries. radio made a very suitable hobby for an impecunious schoolboy. Almost all the bits one used in a circuit could be removed and re-used in a different and improved arrangement. At that time. the adage that it was 'better to travel hopefully than to arrive' was useful to remember.

I also found that there was considerable scope for adding to my pocket money by cross-trading between the various second-hand dealers. especially in battery valves. These were nearly always in working order if their filaments were intact – a fact that could usually be verified by inspection. ■

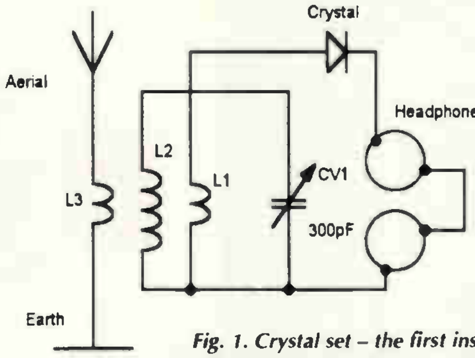


Fig. 1. Crystal set – the first inspiration for many of today's electronics engineers and a device that children still find amazing.

Fig. 2. Finding that Mr Bare's one-valve radio circuit did not work, I ventured into the library where I found a host of books that not only showed me what the error in the circuit was, but also that there were better designs to be had.

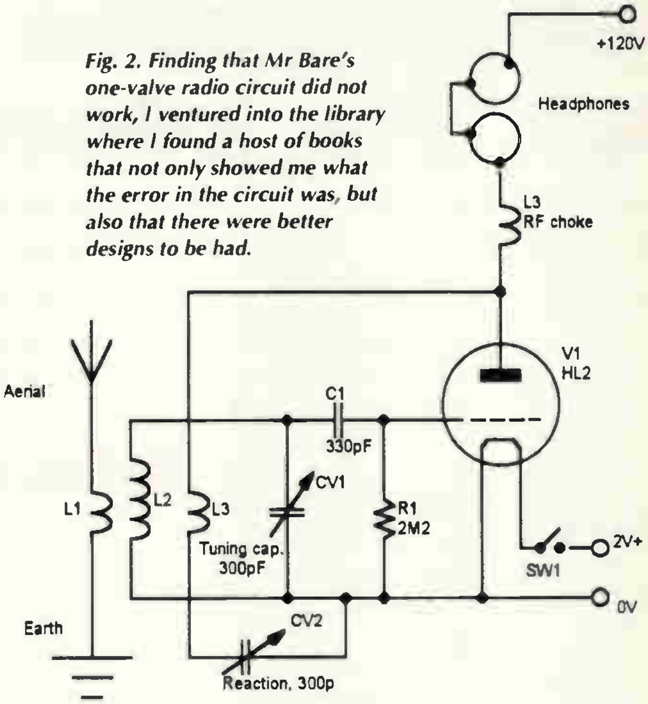
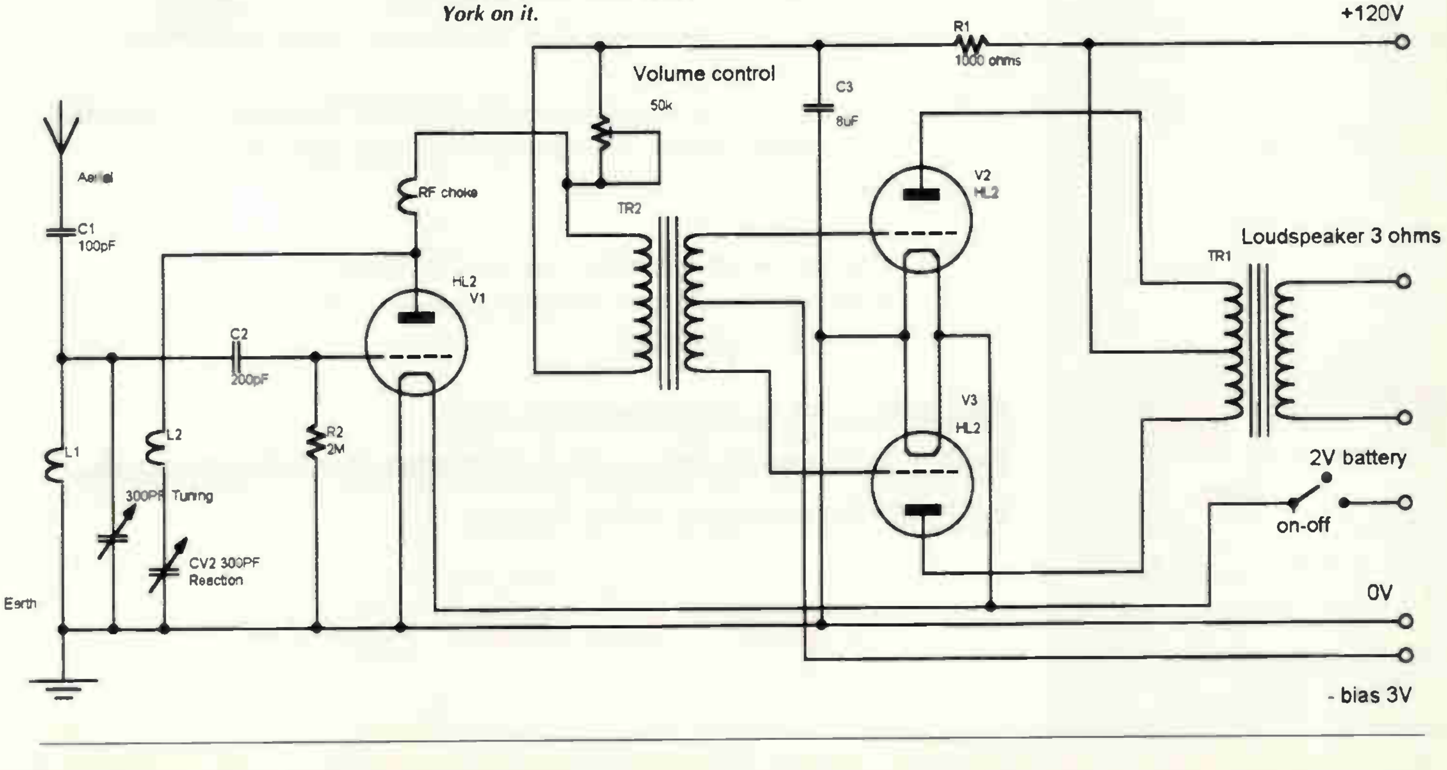


Fig. 3. My first three-valve receiver. Among other things, I used to listen to news broadcasts from RCA in New York on it.



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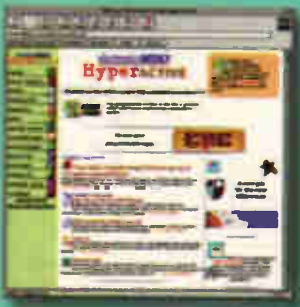

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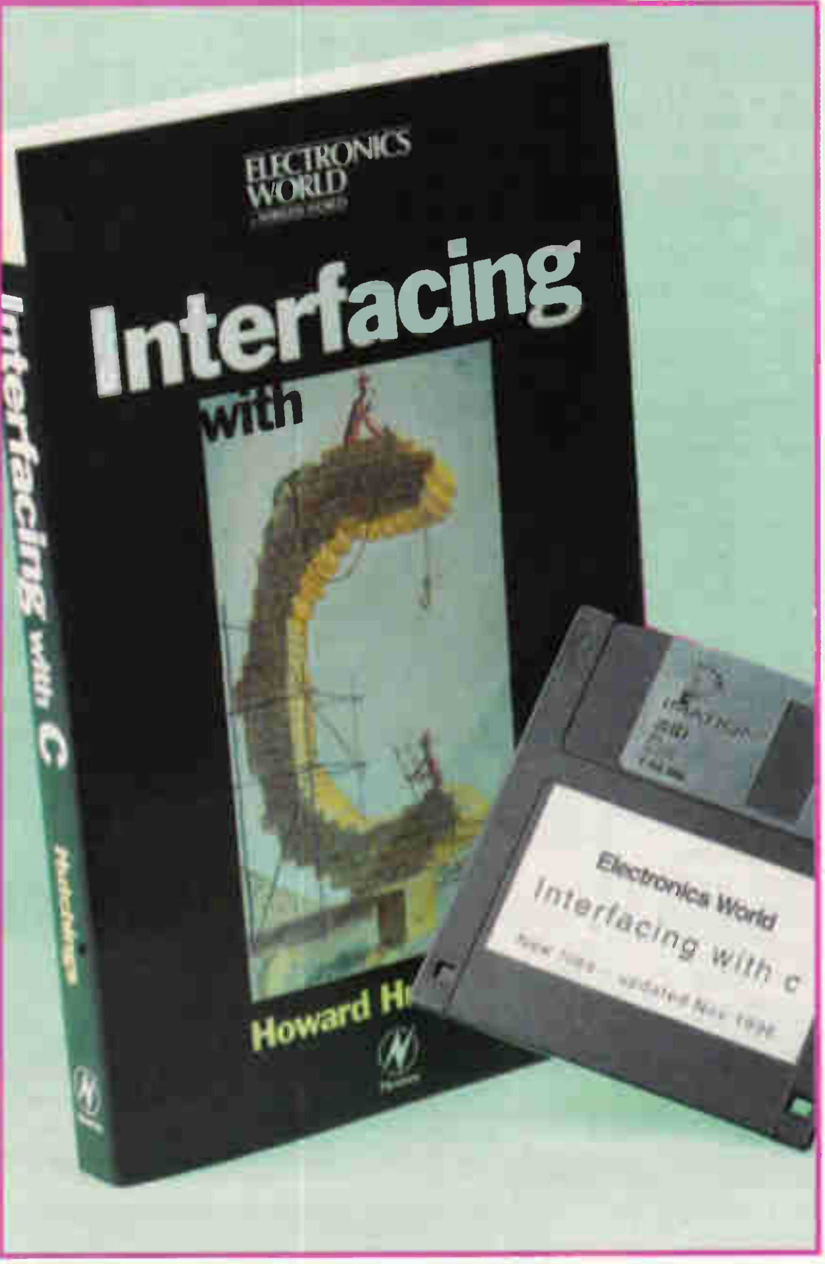

Without an engineering degree, a pile of money, or an infinite amount of time, the revised 289-page Interfacing with C is worth serious consideration by anyone interested in controlling equipment via the PC. Featuring extra chapters on Z transforms, audio processing and standard programming structures, the new Interfacing with C will be especialy useful to students and engineers interested in ports, transducer interfacing, analogue-to-digital conversion, convolution, digital filters, Fourier transforms and Kalman filtering. Full of tried and tested interfacing routines.
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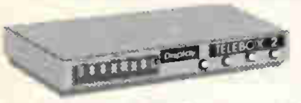

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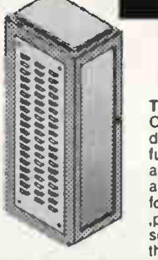

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Astec
Tel: 01384 842211
Enquiry No 501

Crosspoint switches

Vitesse has launched 32 x 32 and 64 x 64 crosspoint switches capable of speeds up to 1.6Gbit/s. The VSC851 and VSC852 are for Gigabit Ethernet and ATM systems that require 1.25Gbit/s or high definition TV that requires 1.5Gbit/s. They eliminate the need for resynchronisation and driver ICs. The VSC851 has 1.6Gbit/s fully differential serial i/os. It has a non-blocking architecture with standard TTL interfaces for control logic and ECL interfaces for high speed differential serial i/o. It has less than 150ps duty cycle distortion and less than 250ps skew in broadcast mode. The VSC851 comes in a 256-pin LDCC package and the VSC852 in a 440-pin BGA.
Vitesse
Tel: 01634 683393
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GSI Technology is sampling 0.25μm GS84018 and 32136, 4Mbyte synchronous SRAMs. They operate at up to 200MHz to provide access times down to 5ns in pipelined mode and 7.5ns in flow-through mode. Choice of mode is user-configurable and users can select linear or interleaved burst operation. The 3.3V devices come in 256k x 18, 128k x 32 or 128k x 36 organisations and have power consumptions up to 360mA for the 200MHz part under worst case operating conditions. Packaging options are 100-lead TQFP or 119 bump BGA and they come in commercial (0 to +70°C) and industrial (–40 to +85°C) types.

Vitesse
Tel: 01332 765983
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Telcom Semiconductor's TC56 low supply current (13μA typ. at V_{out}=3V), low-dropout CMOS linear regulator offers a 10V maximum input voltage range. According to the supplier, the CMOS construction eliminates wasted ground currents typical of bipolar regulators for greater system efficiencies, and longer operating time in battery-powered systems. The device enters shutdown mode when the shutdown control input is low. During shutdown, the regulator is shut off, and supply current falls to 0.5μA typical. Normal operation is restored when the control input is returned to a logic high. The TC56 is available in factory-programmed output voltage settings between 2.1V and 5V in 100mV increments. It is available in a SOT23-5 package.
Telcom Semiconductor
Tel: 0802 770118
Enquiry No 506

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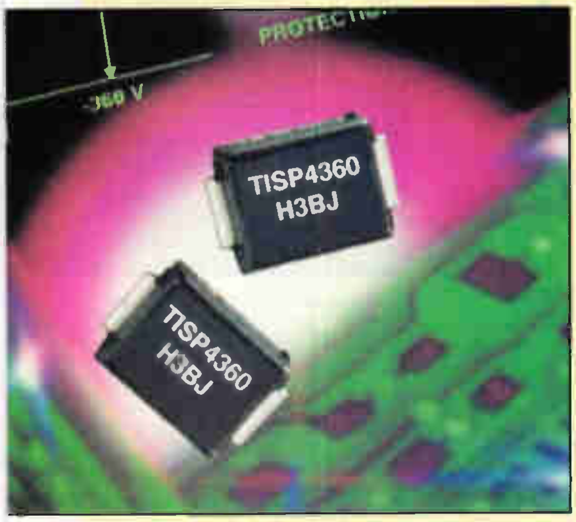
A Fujitsu CAN microcontroller for automotive and industrial embedded control applications is fabricated in 0.5μm CMOS. The MB90F497 uses the firm's F2MC-LX 16-bit microprocessor core and CAN bus macro. A Full-CAN bus interface, conforming to V2.0 part A and B, is combined with onboard 64kbyte single voltage flash as user programmable memory. The message buffer scheme uses eight buffers. A system clock employing an on-chip PLL clock multiplier provides an internal 62.5ns instruction cycle time from an external 4MHz clock. A 32kHz subsystem clock is also available for power saving modes and real time measurement. Power saving modes include sleep, stop, CPU intermittent mode and hardware standby pin.
Fujitsu Microelectronics
Tel: 01332 765983
Enquiry No 505

Multiphase power controllers

A family of advanced multiphase controllers introduced by Intersil is designed to meet the demanding needs of next-generation central processing unit (CPU) core power requirements. The chip-set consists of main controllers (HIP6301, HIP6302 and HIP6303) and FET drivers (HIP6601, HIP6602 and HIP6603), and it features a phase-to-phase current balance and reference voltage tolerance which is designed to provide two to four phase architecture and optimising converter designs for 35 to over 100A operation. The chipset's channel-to-channel current balance scheme has a closed loop individual channel feedback architecture. This scheme allows the motherboard designer to avoid over-designing power components, often necessary when less capable balance



ADSL modem surge protector

A surge protector from Power Innovations is designed to protect the latest ADSL modems and splitters from lightning and power line induced surges. The addition of the ADSL ±15V digital signal on top of the traditional analogue telephone 269V peak waveform means that previous protectors rated at 275V will clip ADSL signals. The firm's high voltage bipolar silicon process has enabled the development of the two terminal, surface mount TISP4360H3BJ with a rated working voltage of 290V to provide close protection without clipping of ADSL signals. The high surge ratings of the DO-214AA packaged SMT protector are complemented by high AC current ratings, simplifying fuse co-ordination and UL1950 compliance. The device offers an off-state leakage of less than 10μA and 24pF of capacitance at 50V, which minimises the loading of the line. It also includes full bidirectional, voltage triggered protection which will clamp surges of either polarity to a low voltage.
Power Innovations
Tel: 01234 223022
Enquiry No 502

methods such as precision duty cycle matching are implemented. Intersil's proprietary closed loop active current balance implementation results in channel to channel current balance of better than ten per cent.
Intersil
Tel: 01344 350250
Enquiry No 507

Single-output 600W supply

Lambda has extended its JWS family of switched-mode power supplies with a single output 600W unit. The JWS600 has been designed to meet the highest requirements for power and reliability in process control, industrial ATE, telecoms, materials handling and industrial computing applications. With universal input values of 85-265V AC or 120-330V DC, users have the choice of seven models providing output voltages from 2V up to 48V. Built-in power factor correction ensures that power factor rates are typically 0.99 at 100V AC full load and 0.95 at 200V AC. Over current, over voltage and thermal protection are all provided. The fan-cooled units measure 190 x 92 x 200mm and can be operated at temperatures between –10°C and +60°C.

Lambda
Tel: 01271 856600
Enquiry No 508

VDSL filters and transformers

Pulse has announced the first set of filter modules and hybrid transformers that meet the FSAN committee recommendations for VDSL broadband access. The solution was designed in partnership with Savan Communication, a leader in VDSL, and Infineon Technologies, a worldwide leader in semiconductors for xDSL technologies. The new components, which are part numbered B4010 to B4018, encompass all of the magnetics required from the twisted pair connector to the Infineon/Savan VDSL analogue front-end. VDSL technology provides symmetric and asymmetric data rates, thereby enabling local and long-distance telephone companies to provide broadband services over existing copper lines. The joint Infineon/Savan VDSL chip-set and Pulse's set of filters will enable rates of up to 30Mbit/s. The devices include a POTS and ISDN splitter for VDSL, transmit and receive filters and an integrated hybrid and band pass filter. Two versions are available, one for 100Ω lines and one for 135Ω lines. This flexibility provides the optimal performance for different countries and applications.

Pulse
Tel: 01483 401700
Enquiry No 510

Mini frequency converters

Atlantic Microwave has announced a range of miniature frequency converters for use in microwave communications links. With frequency coverage from 6 to 16GHz, the AFC series of miniature frequency converters are designed for use in terrestrial microwave links and are believed by the supplier to be one of the smallest integrated assemblies currently available for such applications in broadcast and related communications. At the heart of each unit is a dielectric resonator oscillator using a GaAs FET as the active device and with the output buffered by a further GaAs FET gain stage. This microwave signal feeds into the local oscillator port of a microstrip double balanced mixer which can act as either an upconverter or down converter with inverting or non-inverting IF. Individual units can be mechanically tuned over ±30MHz and, once set, maintain a stability of frequency within ±120ppm over a temperature range of –10 to +60°C. Conversion loss from RF to IF is typically 7dB and this is flat over the DC to 4GHZ IF range to within 1dB.

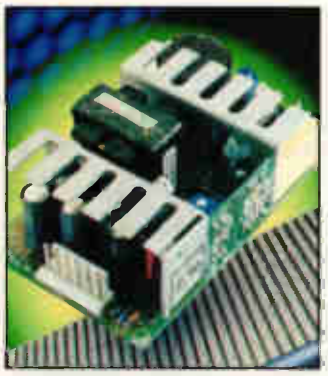
Atlantic Microwave
Tel: 01376 550220
Enquiry No 511

Supervisory circuit enables DSP monitoring down to 1.2V

A family of supervisory circuits from Texas Instruments is intended to enable designers to monitor DSP and microprocessor core voltages down to 1.2V. The family of processor supervisors offers an integrated watchdog and manual reset function in addition to the ability to monitor supply voltages as low as 1.08V. The family of supervisory circuits, the TPS312x, provides precision initialisation and timing supervision of DSPs and other processors operating at 1.2, 1.5, 1.8 and 3V. The devices have an integrated watchdog timer that monitors processor activity and a manual reset function. In addition, the TPS312x family has an active-low and active-high reset function that enables designers to use these devices in a wide range of DSP and processor-based systems. All the devices in the family only need a minimum supply voltage of 0.75V, and the integrated Power-On reset function is enabled as soon as the supply voltage rises above the 0.75V level.
Texas Instruments
Tel: 01604 663399
Enquiry No 509

40-110W power supplies

Condor DC Power Supplies has a range of power supplies with output power varying from 40 to 110W which is available in both commercial and medical styles meeting EN60601. Both styles are available in single, dual and quad outputs with output combinations of 5V, 12V, 15V, 24V and 28V DC in 40W, 50W, 65W, 75W and 110W models. Medical range models are also additionally available in 140 and 150W package styles. Input ranges from 90 to 264V AC. Some products achieve power densities of over 4W/in^3.



Condor DC power supplies
Tel: 01769 540744
Enquiry No 512

10V linear supply

Calex Electronics' latest linear power supply, the CMI.10.400 provides a regulated 10V output at 400mA. Input options are 100,110, 220, 230 and 240V AC. The supply is capable of driving eight 350Ω bridge load cells as would typically be found on a truck weighing platform. The output voltage accuracy ±1 per cent. Line and load regulation is ±0.1 per cent. Noise and ripple is 1mV RMS. Input to output isolation is 1500V RMS. The isolation resistance is 50MΩ. The input frequency range is 50-60Hz.
Calex
Tel: 01525 373178
Enquiry No 513

Magnetic circuit breakers

The Airpax IEGZX range of magnetic circuit breakers being stocked by EAO, provide power switching, circuit protection, and circuit control over the current range 0.05-50A at maximum voltage ratings of 80V DC or 250V AC (50/60Hz or 400Hz). In addition, the devices meet the key IEC spacing requirements demanded by international equipment safety standards including VDE, UL and CSA.

EAO
Tel: 01444 236000
Enquiry No 514

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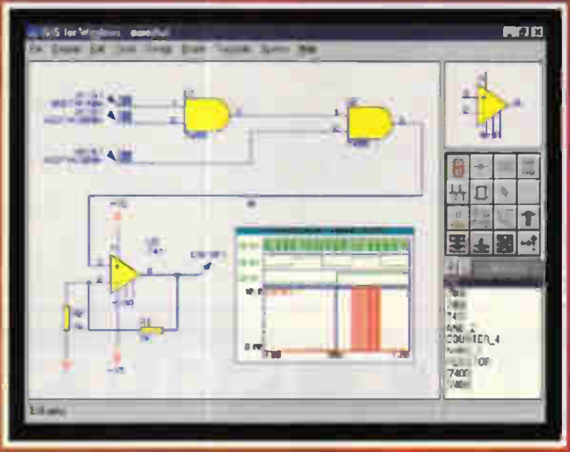

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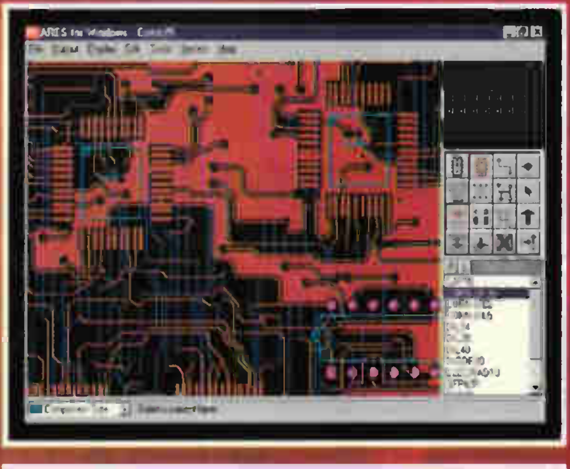

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CIRCLE NO. 122 ON REPLY CARD

Optical fibre testers for LAN

Microtest has designed a family of fibre optic test tools that measure the optical loss of single and multi-mode fibre optic cables and components at various wavelengths. Called SimpliFiber it features data-storage, auto-wavelength detection and the company's ScanLink PC interface. Standard kits include a dual wavelength 850/1300nm LED light source and an 850/1300/1310/1550nm power meter as well as ScanLink software and a serial cable for easy PC uploads of measurement data. An optional 1310nm laser source is also available and perfect for testing single-mode fibre cabling. Optical sources are stable and repeatable for consistently accurate results, and auto-wavelength detection and data recording minimise common testing errors.
Microtest
Tel: 01293 894000
Enquiry No 515

Triple channel 8-bit a-to-d converter

E2000 is offering the ICS1531, a triple-channel 8-bit a-to-d converter with an integrated line-locked clock generator. Using ICS's advanced low-voltage CMOS mixed-signal technology, the device is designed for data-capture at resolutions from VGA to UXGA. It offers A/D data conversion and synchronised pixel clock generation up to 160Msamples/s or 160MHz, in a single-ended format. Dynamic phase adjust (DPA) circuitry allows end-user control over the pixel clock phase, relative to the recovered sync signal and analogue pixel data. Either the internal pixel clock can be used as a capture clock input to the A/D converters or external clock input can be used. It provides either one or two pixels per clock.
Electronics 2000
Tel: 01494 444044
Enquiry No 516

High-speed, low-power amps

Burr-Brown's INA133 and INA143 families of high speed, precision difference amplifiers is available through Silicon Concepts. Both families are available in single and dual versions. The INA133 (G=1) includes wide bandwidth (1.5MHz), high slew rate (5V/µs), fast settling time (5µs to 0.01 per cent), low offset voltage (450µV max), low offset drift (5µV/°C max), and low gain error. The 1NA143 provides gain of 10 or 0.1 and 150kHz (G=10) with similar specifications.
Silicon Concepts
Tel: 01428 751617
Enquiry No 517

LCD flat panel reads in sunlight

A sunlight readable colour LCD XGA 38cm TFT flat panel has been launched by Digital View and 26.4cm VGA and 30.7cm SVGA panels are planned. The DView DV-7000 is built around a base Sharp TFT panel with additional backlights and replacement light filters. The backlight uses 12 CCFL tubes and an inverter to provide 1500cd/m^2 screen brightness and a colour spectrum close to the colour temperature of typical sunlight (5000K). Consuming 35W at full power, the backlight can be driven at intensities from 10 to 100 per cent. The inverter can be dimmed using pulse width modulation and, with an optional light sensor, can dim automatically as ambient light decreases. Integral interface controllers provide analogue PC and video (Pal, NTSC and Secam) input or PCI and LVDS connectivity. They are made from aluminium, alochromed with a protective powder finish. The front bezel can be used to seal the front panel to IP64, and there is an optional membrane keypad to allow external control of the OSD. Standard controls include on-off, brightness, contrast and auto-set up, with an option of infra-red control of OSD. All are Vesa DPMS power management and Vesa DDC2/b plug-and-play compatible, and require a 12V DC power supply.
Digital View
Tel: 0208 236 1112
Enquiry No 526

Open frame AC-DC supplies

Acal has introduced the VLT-130 range of 18 quad and single-output, open-frame AC-DC power supplies. Made by EOS, they accept inputs from 90 to 264V AC and output up to 130W with forced-air cooling or 100W with convection cooling. Measuring 7.6 by 12.7 by 2.5cm, the supplies have a typical efficiency at full load of 82 per cent. They weigh less than 300g. The quad-output supplies can have 3.3, 5, 12, –12, 15, –15 or 24V DC outputs. The 3.3V output unit offers up to 16A and the 5V up to 14A. The 4100 versions come with single wire current sharing across V1 and V2 outputs allowing a combined 30A. This makes them suitable for networking products such as LAN switches, hubs, routers, edge servers, access concentrators and multiplexers. The single-output models have 2.5, 3.3, 5, 12, 15, 24 and 48V DC outputs. They can be used as stand-alone power systems for chipsets that require 2.5 and 3.3V power and current up to 40A and for distributed products using 24 and 48V. The 1100 versions come with single wire current share for redundancy or n+1 applications. Power factor correction meets EMC standards, including IEC1000-3-2, -4-2, -4-4 and -4-5. They have overvoltage, overload and short-circuit protection.
Acal Electronics
Tel: 01252 858585
Enquiry No 518

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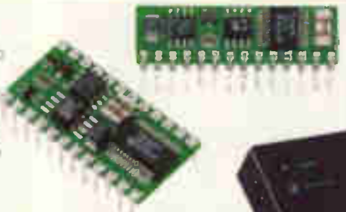

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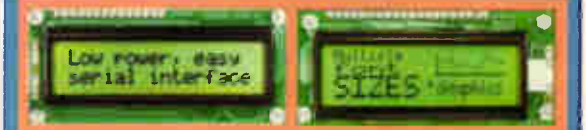

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Toshiba has added QFPs to its range of lead-free packages for Asics. The 64, 80, 100, 120, 144, 160 and 208-pin devices use palladium plated frames combined with a mould resin. The lead-free chips range from 7.5mm^2 60-pin to 12mm^2 208-pin QFPs that can withstand lead-free reflow soldering processes. Future products may use tin-silver plating and solders.

Toshiba Electronics
Tel: 0208 938 4644
Enquiry No 519

Hi-fi terminal plates

Cliff Electronic Components has introduced hi-fi terminal plates for installation on top, middle and entry level speaker cabinets. Made from airtight polypropylene, the crossover can be rear mounted vertically or horizontally with no top-hole option. Terminals are gold plated solid brass rated at 30A minimum. Gold plated links are available. Terminals are fitted flush to the surface and a rear tag allows PC mounting or a 0.63cm push on receptacle.

Cliff Electronic Components
Tel: 01737 771375
Enquiry No 520

Card readers

Omron has introduced a smartcard reader that provides PC security and secure transactions over the Internet. For OEM installation, the V4IF reader fits into the standard floppy drive bay of a desktop PC. The unit is manually operated and compatible with ISO7816 smartcards. There are various T=0 and T=1 processor cards available. Communications with the host PC uses RS232C protocol via a serial i/o card, which also provides the +5V DC power supply to the reader. The unit sets data transmission speed automatically in synchronisation with the host, offering a choice of 1200, 2400, 4800, 9600, 19,200 or 38,400bit/s across the serial link.

Omron
Tel: 0208 450 4646
Enquiry No 521

Outdoor enclosures

Optima's street-and-track side cabinets are for cable TV, telecoms, rail communications, security and protection applications. Users are not restricted to set sizes; they are built to suit the application. They can be supplied with vandal resistant locks and hinges, combined RFI and EMI protection and environmental protection exceeding IP55.

Optima Enclosures
Tel: 01875 610747
Enquiry No 522

Linecard IC

Infineon's Delic-LC PEB20570 and Delic-PB PEB20571 linecard controller ICs are for use in PBXs and central office systems and central controllers in small and mid-size PBXs. The two chips with the PEB20590 VIP ISDN transceiver are available on an evaluation board, the SIPB20570. The LC version provides switching, multiple HDLC and layer one control for up to three VIPs (24 ISDN ports), connected via the IOM-2/GCI interface. Other transceiver ICs may be connected via IOM-2/GCI ports, letting it provide up to 16 ISDN channels or 32 analogue lines. Equipped with four PCM ports, access to data highways running up to 8.192Mbit/s is possible. On-chip features include a switching matrix of 160 by 128 TS, serial communication controller operating up to 16.384Mbit/s and A-law to µ-law conversion.

Infineon
Tel: 01344 396315
Enquiry No 523

Comms board

Radstone has introduced the Empower EPIA-8240 Cots VMEbus communications board in 6U and PMC form factor. For embedded applications, it is based on the MPC8240 from Motorola, which combines a PowerPC 603e microprocessor core, memory controller, PCI bridge and other logic functions. The board has two PMC slots, 64Mbyte DRAM, flash capacity, 100baseT Ethernet support and a compact flash site. Onboard build options include four sync and async serial communication channels implemented by an MC68360 Quicc and a MIL1553 interface for embedded communications or control applications. In addition, PCI subsystem expansion is possible for air and conduction cooled applications, based on PCI mezzanine cards, including MIL1553, ATM, serial, graphics, flash, SRAM

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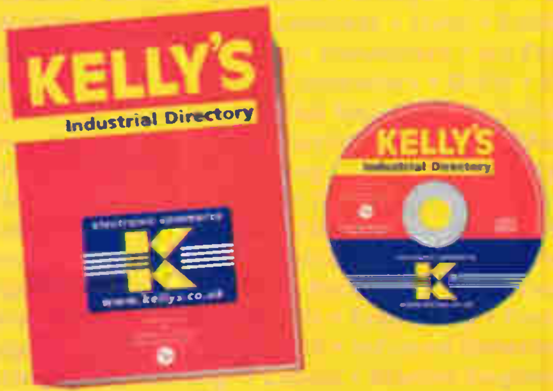

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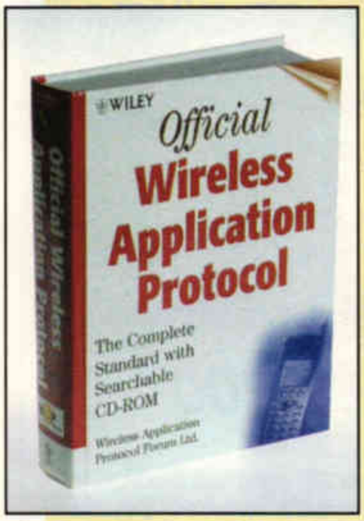
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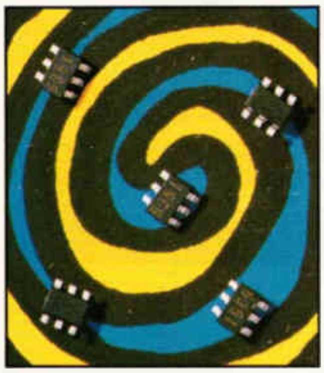
Radstone
Tel: 01327 359444
Enquiry No 524

Voltage-inverter IC pumps iron

A voltage inverter IC from Flint is based on charge pump technology. Using two external capacitors on a standard circuit board, the Torex XC6351 lets negative voltages be produced such as those required for GaAs bias power supplies and negative-power-supply op-amps. Operating voltage is 1.2 to 5.0V and there is a choice of oscillator frequency. With output currents up to 10mA, power requirements are typically 100μA for 35kHz versions and 40μA for 12kHz. Transition efficiency is 95 per cent. There is a chip-enable function to halt IC operations and power down the output, lowering requirements to 2μA standby. It comes in six-pin SOT-26 packaging. Applications include palmtop computers, handheld organisers, cellular and portable phones, and miniature LCD panels.
Flint Distribution
Tel: 01530 510333
Enquiry No 525



16-bit micro

Hitachi Europe has announced the 16-bit H8S/2345 microcontroller for industrial and mass market applications. It is available in a ROM-less version, the H8S/2340, and a 128kbyte flash version, the H8S/2345F, both with 4kbyte RAM. It comes in a 100-pin package. The CPU provides about 6Mips Dhrystone running at 20MHz, 5V, and executes basic instructions in one cycle. A 1:1 crystal oscillation helps reduce unwanted radiation. The peripheral set comprises a TPU timer unit (6 x 16-bit, with up to 16 IC/OC), two additional 8-bit timers, watchdog, pseudo-DMA, two Usarts, eight-channel 10-bit a-to-d, two-channel d-to-a and bus controller. In single-chip operation, up to 71 i/o pins and eight inputs are available, some of which offer 10mA drive, programmable open-drain, pull-ups, and Schmitt-triggered IRQ capability.
Hitachi
Tel: 01628 585163
Enquiry No 527

Linear motion guides

Unimatic has introduced the SBC linear motion guides for use in CNC machines and automated production lines. Four ranges are available. FL and SL types come in eight sizes each with block heights from 24 to 90mm. FLL and SLL versions come in seven sizes with block heights from 30 to 90mm. SL and SLL types have narrower bodies than FL and FLL versions, which can be attached from the top or bottom of the carriage. Rails can be supplied to length. They have a circular arc contact construction. This means the balls contact in two areas letting the linear motion guides absorb the tolerances caused by elasticity. Ball bearings in the guide block are protected with a seal while lubrication can be provided via a grease nipple. Accessories include a metal scraper to remove debris from the rails as the block moves up and down, side grease nipple for where the standard position is obstructed or inconvenient, bellows to prevent ingress of dust and debris, and a hole cap to prevent pollution inside the bearing.
Unimatic
Tel: 0208 922 1001
Enquiry No 529

Chip cooler

Fan Technology has added to the Adda AP series of chip coolers. The Neptune AP5008 has a frame size of 50 by 50mm and a profile of 8mm. It has air outlets on all four sides of the fan frame. The eight-pole motor operates from a 5V supply, is available in three speed options and has a power rating of 0.5W.
Fan Technology
Tel: 01403 275131
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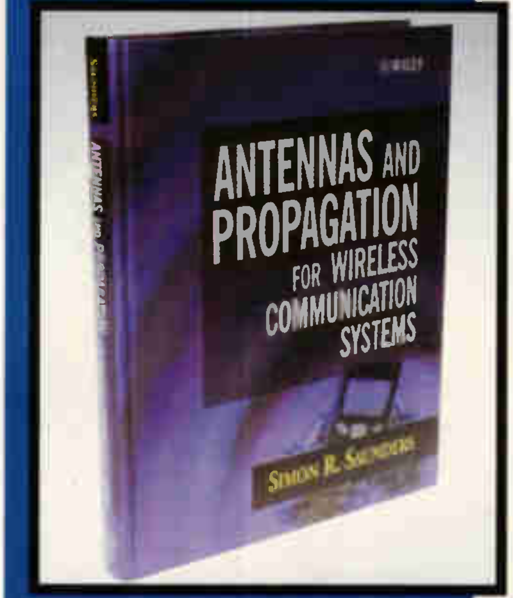

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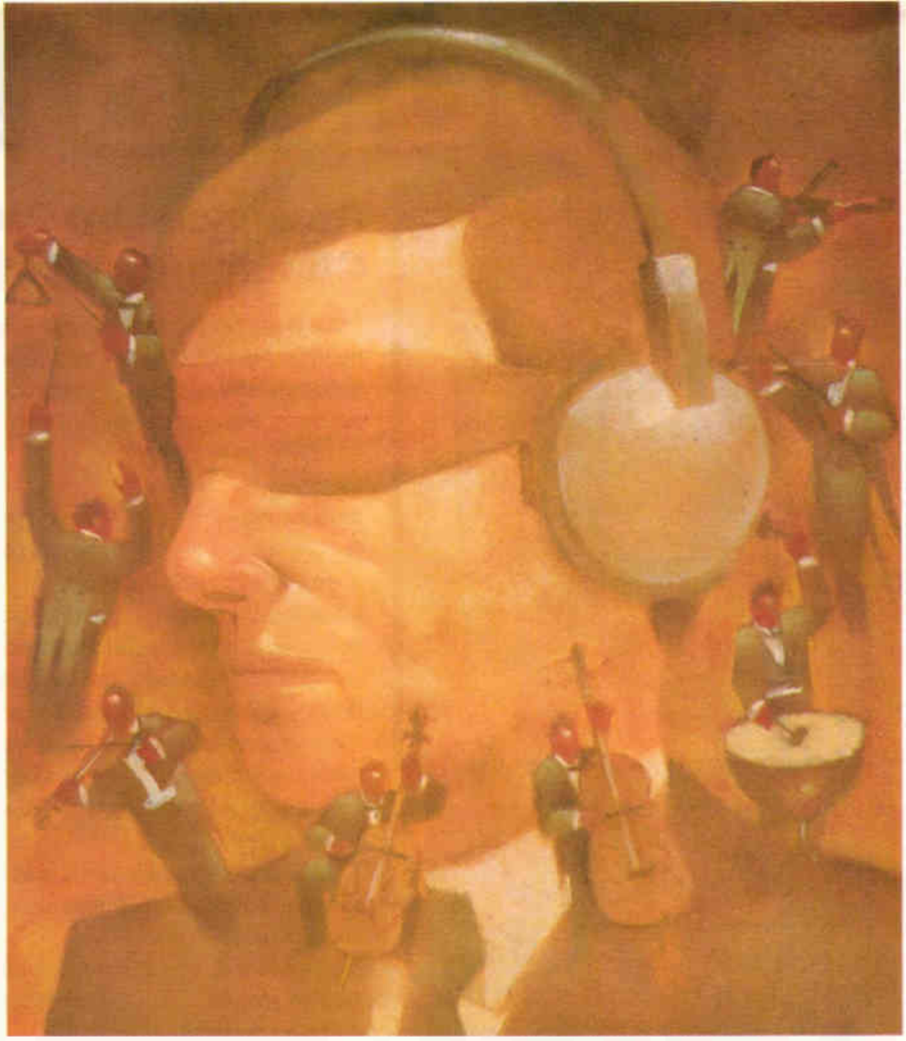


Stereo from all angles V

Loudspeakers are the most difficult aspect of stereophonic reproduction. While existing microphones are not ideal, they have reached a standard that cannot be criticised by listening through today's loudspeakers. John Watkinson explains why stereo requires more accurate speakers and presents a new measurement technique.

It is axiomatic that the accuracy of the stereophonic imaging system needs to be at least as good as the accuracy of the human ear, or deficiencies will be heard. A complete knowledge of the human auditory system will allow suitable quality criteria to be set down so that systems can be designed to a suitable performance.

An earlier part of this article set showed that the ear is a lossy device because it exhibits masking. Not all of the presented sound is sensed.

If a lossy loudspeaker were designed to a high standard, the losses may be contained to areas that are masked by the ear, and then that loudspeaker would be judged transparent. Douglas Self has introduced the term 'blameless' for a device whose imperfections are undetectable.

This is good news for the loudspeaker designer because the ear has finite accuracy in frequency, time and spatial domains. This means that a blameless loudspeaker is not just a concept. It could be made real by the application of sufficient rigour.

Although sufficiently complete knowledge of the human auditory system exists, the loudspeaker industry has, with a few exceptions, failed to act on it. Instead, it delivers commoditised products at low cost with a correspondingly low performance that has not improved in years.

Audible defects are introduced into the reproduced sound in frequency, time and spatial domains, giving the loudspeaker a kind of character which is best described as a signature or footprint.

The stereophonic sound field at the listener carries information in frequency, time and spatial domains. Unless sufficient accuracy exists in each domain there will be a loss of realism. While loudspeakers with an adequate frequency response are relatively easy to find, a sufficiently accurate time and spatial response is much more elusive.

The legacy loudspeaker is designed solely to have a flat on-axis frequency response, because traditionally that was all that was thought necessary. The time and spatial domains were neglected almost totally.

A fixation with the frequency domain

A scant knowledge of communications theory will reveal that a sine wave has no bandwidth and carries no information, and this must also be true if that sinewave is carried as sound. In fact the majority of information in sound is carried in transients, and a fixation with the frequency domain is of no help in getting the transient response accurate enough.

What is needed is a balanced design where frequency, time and spatial characteristics are afforded sensible proportions of design effort. The critical domains will be considered briefly.

The requirement for linear addition of an unlimited number of signals puts a stringent requirement on the distortion characteristics of the whole system from microphone to speakers. Non-linearity in a mono system will result in intermodulation products, but some of these may be masked by the signal because signal and harmonics emanate from the same single speaker.

In a stereo system, intermodulation between signals due to sound sources at different locations may cause distortion products at other locations in the virtual image. Attentional selectivity – the 'cocktail party effect' explained in the second of these articles* – allows the ear to detect sounds in different places that would be masked if they were coincident. It follows that the distortion criteria for stereophonic equipment must be more stringent than for mono.

The normal criterion for frequency response in mono is relatively undemanding, but this is also irrelevant for stereo. In the November issue, I illustrated the difference in level needed between speakers to produce a given location of the sound source. That figure is reproduced here as **Fig. 1**.

Listening angles

If any angular accuracy in the image is required, then the function shown in that figure must define the accuracy to which the frequency responses of the two speakers must track. If we want an angular accuracy of, say, 5° the tracking has to be within an eighth of a decibel.

As we are considering geometry, the decibel is not an appropriate unit here and it is better to use linear units. The frequency response tracking tolerance then becomes about 1.5%, which is feasible with care.

In practice the listener will not be exactly on the forward axis of both speakers, and may even subtend a slightly different angle to each. Under these conditions, the directivity of the speakers becomes critical.

The criterion must be that for angles likely to be subtended by the listener, the frequency response must be as accurate as it is on axis. This aspect of speaker design is badly neglected and leads to the requirement that the listener sit in a precise 'sweet spot'. In fact the existence of a small sweet spot is evidence of a poor loudspeaker.

A spatially-accurate virtual image can only be obtained if the loudspeaker acts as a point source. This is because the size of the original sound source is convolved with the acoustic size of the speaker.

Audio and optics

There is a direct analogy here with optics, as both are concerned with image reproduction. **Figure 2a)** shows that in optics the image is reduced in resolution because it is convolved with the spatial impulse response of the lens. Figure **2b)** shows that the acoustic image is convolved with the spatial impulse response of the loudspeakers. In both cases the ideal is an impulse response comprising a singularity or narrow spike. In a lens, this requires infinite aperture; in a speaker this requires zero acoustic size, which is a

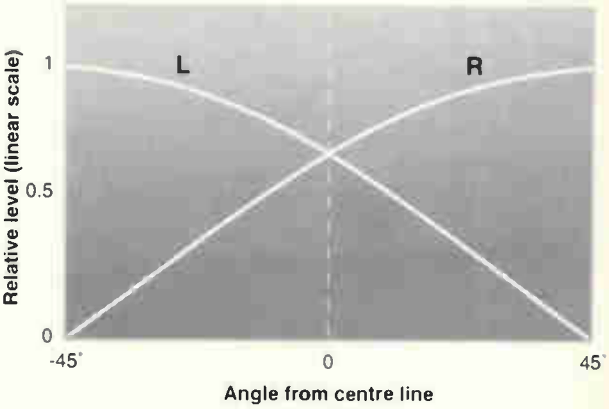
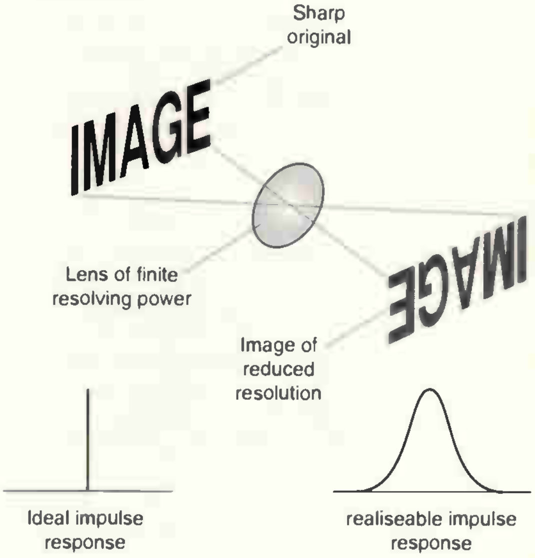


Fig. 1. Apparent position of the virtual sound source is a function of the level difference between the channels. This defines the characteristics that a microphone must have.


Sharp original
IMAGE
Lens of finite resolving power
IMAGE
Image of reduced resolution
Ideal impulse response
realiseable impulse response

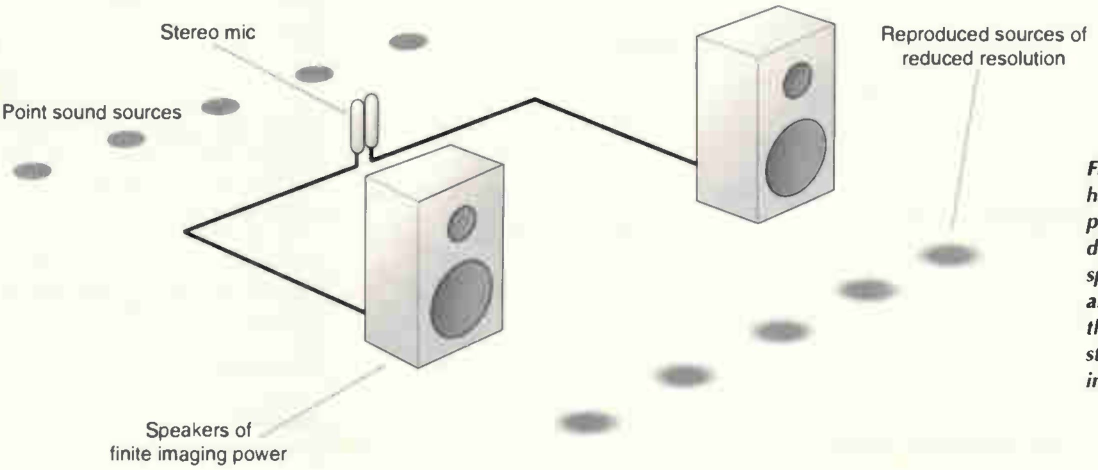

Stereo mic
Point sound sources
Reproduced sources of reduced resolution
Speakers of finite imaging power

Fig. 2a) Optical systems have measurable imaging performance and can be designed for a given specification. In b), although the mechanism is the same, there is no standard unit of speaker imaging accuracy.

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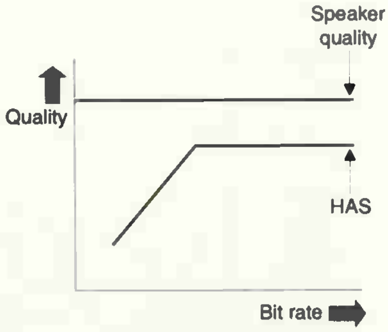
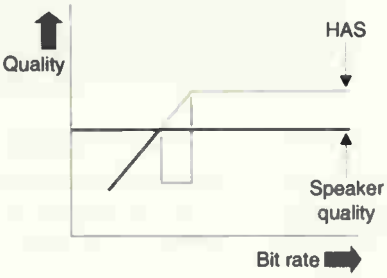
*August 1999

Fig. 3a) Where speaker performance is sufficient, the compressor is tested. Where speaker performance is inadequate, as in b), the codec tests the speaker!



Speaker quality exceeds human auditory system – test is valid



Human auditory system quality exceeds speaker quality so test is invalid.
A lower bit rate will appear adequate: this is the bit rate of the speaker.

point source.

Recently, a new form of flat panel loudspeaker has emerged. These work by chaotic excitation of a panel with bending waves. Clearly such a distributed-mode loudspeaker can never act as a point source, making it unsuitable for accurate stereo reproduction.

Point-source speakers must minimise reflections from their own enclosures. If they don't, the acoustic size of the speaker becomes the enclosure width. Unfortunately, powerful enclosure reflections are exactly what are provided by the legacy rectangular loudspeaker with sharp corners. These reflections are due to acoustic impedance changes.

If you could see the sound, you would double up with mirth at how ineptly it was being radiated.

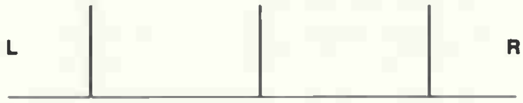
A new type of measurement

Traditional loudspeaker measurements do not measure the spatial accuracy or acoustic size of the speaker. Another technique is needed to do that.

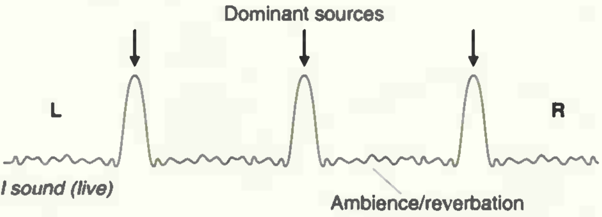
The stereo loudspeaker system can be modelled as an information channel of finite capacity which can actually be measured as an equivalent bit rate. A poor pair of loudspeakers will measure as having a low bit rate.

Psychoacoustic researchers have known for some time that poor quality loudspeakers give erroneous results when measuring hearing loss. Hearing loss is a reduction in information capacity of the ear and trying to measure this with a speaker of reduced capacity is not useful. Loudspeakers with a reputation for realism also give the highest scores when performing intelligibility tests on hearing impaired subjects.

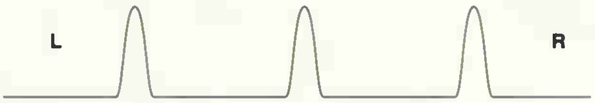
In professional audio, the ability of an engineer to monitor sound quality can only be as good as the information capacity of the speakers used. When the speaker information capacity is limited, the presence of a defect in the signal source may go unheard and it may erroneously be assumed that all is well when in fact it is not.



a) Pan-potted mix



b) Original sound (live)



c) After compression - ambience lost

Fig. 4. In stereo, the result of bit-rate reduction is a loss of ambience.

Compression effects

Audio bit-rate reduction or compression is becoming popular, but it has some undesirable characteristics for stereo. It should follow from what has been said above that compression codecs can only meaningfully be assessed on speakers of adequate information capacity. It also follows that the definition of a high quality speaker is one which readily reveals compression artifacts.

Non-ideal loudspeakers act like compressors in that the distortions, delayed resonances and delayed re-radiation they create conceal or mask information in the original audio signal. If a real compressor is tested with non-ideal loudspeakers, certain deficiencies of the compressor will not be heard. The spatial compression of non-ideal stereo loudspeakers conceals real spatial compression artifacts.

Lossy compression does not preserve the original waveform, but seeks to be blameless by placing the noises where they will be masked. This can be achieved at a high enough bit rate, beyond which no improvement in quality would be observed.

Naturally, one would want to carry out listening tests to see if this goal had been achieved. If blameless loudspeakers are used, the test is valid as shown in **Fig. 3a)**. As the bit rate increases, the quality levels off where the human auditory system is masking all of the compression artifacts.

However, the legacy loudspeaker is not blameless. When a loudspeaker has a signature, how are we to know that the speaker signature is not masking compression artifacts? When listening in series, as we must, on hearing a deficiency, how are we to determine whether this was the codec or the speaker?

Bit-rate evaluation

Figure **3b)** shows that if the test of Fig. 3a) is repeated with a legacy speaker, the bit rate at which the quality levels off is artificially reduced because the speaker is masking the compression artifacts with artifacts of its own. In this case we have measured the bit rate of the speaker.

All that is needed is a variable bit-rate audio codec. Clearly with a very high bit rate, the speakers will be the limiting factor, whereas with a low bit rate, the codec will dominate. At some intermediate bit rate, the effects will be equal. At this point the masking due to the speaker is equal to the level of artifacts from the codec.

At any lower bit rate, compression artifacts will become audible over the footprint of the speaker. The worse the information capacity of the speaker, the lower the bit rate at which the artifacts are audible.

As a result by simply varying the bit rate of a codec, it becomes possible to measure the effective bit rate of a pair of loudspeakers. The measured bit rate reflects the spatial or imaging accuracy in a way that other tests do not.

A virtual sound source from a panpot has zero width and on blameless speakers would appear as a virtual point source as in **Fig. 4a).** As a result stereo reverb is added to panpotted mixes and this is audible between the point sources as in Fig. **4b)**. A similar result is also obtained with real sources using a coincident pair of mikes. In this case the sources are the real sources and the sound between is reverb/ambience.

Figure **4c)** shows what happens when accurate speakers are used to assess some audio compressors. Even at high bit rates, corresponding to the smallest amount of compression, it is obvious that there is a difference between the original and the compressed result.

An absence of ambience and reverb

The dominant sound sources are reproduced fairly accurately, but what is most striking is that the ambience and reverb between is virtually absent, making the decoded sound much drier than the original.

However, upon reproducing such a stereo signal with the legacy square box speaker, the point sources have been spread by the speaker footprint so that there are almost no gaps between them, effectively masking the ambience as in **Fig. 5a)**. This represents a lack of spatial fidelity.

Figure **5b)** shows what happens when legacy speakers are used to assess a compression system. The spatial smear increases the dominant source size to such an extent that the ambience between is inaudible. As a result if this ambience is removed by a compressor its loss will not be noticed. This is why poor speakers cannot be used to assess compressors.

MPEG and Dolby

The same effect is apparent to the same extent with both MPEG layer 2 and Dolby AC-2 coders, even though their internal workings are quite different. In retrospect this is not surprising because both are probably based on the same psychoacoustic masking model.

MPEG layer 3 fared even worse because the bit rate is lower. Transient material has a peculiar effect whereby the ambience would come and go according to the entropy of the dominant source. A percussive note would narrow the sound stage and appear dry but afterwards the reverb level would come back up.

If an opportunity arises to compare the same commercially available recording on CD and MiniDisc with accurate loudspeakers it will be obvious that the MD version is inferior. All of these effects largely disappear when the signals to the speakers are added to make mono which removes the ear's ability to discriminate spatially.

The value of subjective testing

Because of the phenomena described here, audio codecs have reached the market which produce audible artifacts even at high bit rates, despite exhaustive subjective testing. When one examines the results of any subjective compression test, it becomes clear that the type of loudspeakers used would have been those having the shortcomings mentioned above. As a result these subjective tests are invalid because the masking of the legacy speakers was masking the coder being tested.

While lossy compression may be adequate to deliver post-produced audio to a consumer with mediocre loudspeakers, these results underline that it has no place in quality stereo reproduction environment.

When assessing codecs, loudspeakers having poor diffraction design will conceal artifacts. When mixing for a compressed delivery system, it will be necessary to include the codec in the monitor feeds so that the results can be compensated. Where high quality stereo is required, either full bit rate PCM or lossless (packing) techniques must be used.

Audio codecs can only be developed fully if blameless monitoring loudspeakers are available. Such precision loudspeakers require no more than an appropriate degree of rigour during the design stage, along with some high-grade circuit design. But they have the advantage that there usually needs to be very little change between the prototype and the production phase.

Interesting conclusions

The ability to measure loudspeaker information rate allows different designs to be compared objectively. This leads to some interesting conclusions.

The measured bit rate of legacy loudspeakers is disturbingly low at about one tenth the bit rate of a CD. This means that 90% of the disc data is discarded in the speaker. Clearly if this is the case any further increase in recording format performance, such as increasing the sampling rate, must be a complete waste of time until attention is given to loudspeakers.

It is also interesting to be able to measure the noise floor of loudspeaker drive units. Most cheap drive units use ferrite magnets. As ferrite is a ceramic, it is an insulator and the magnetic field is free to move within the magnet due to the coil reaction.

Unfortunately this domain-wall movement, which was described by Barkhausen, is not linear. The result is a form of quantised flux modulation that causes audible program modulated noise on the audio. As far as I can see, it is impossible to obtain 16-bit resolution with a ferrite magnet. Conductive magnetic materials such as Neodymium and the older alnico and alcomax do not suffer from this problem. ■

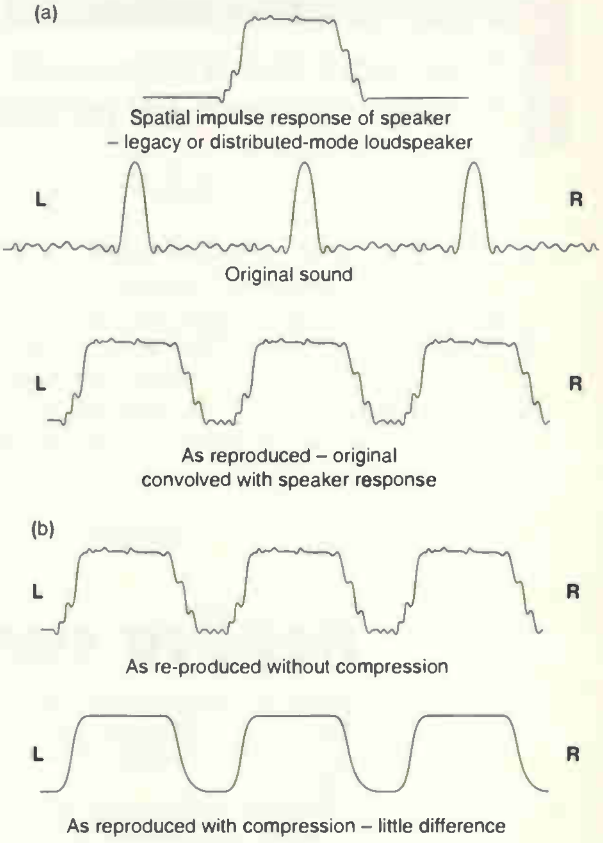

(a)
Spatial impulse response of speaker – legacy or distributed-mode loudspeaker
L
R
Original sound
L
R
As reproduced – original convolved with speaker response
(b)
L
R
As re-produced without compression
L
R
As reproduced with compression – little difference

Fig. 5. In a), point sources have been spread by the speaker footprint so that there are almost no gaps between them, resulting in a lack of spatial fidelity. In b), legacy speakers are used to assess a compression system. Spatial smear increases the dominant source size to such an extent that the ambience between is inaudible.

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Replica correlation detection

In this idea, weak signal reception is assisted by correlated reception of a binary phase-shift keying, or BPSK, coded signal.

The period of the pulse being transmitted is divided into seven cells each of width Δt, and a BPSK coding scheme employed as in **Fig. 1**. Each cell covers an integral number of cycles *n* of the carrier or subcarrier, where *n* is 2 in the case shown.

The receiver includes the replica correlation circuit of **Fig. 2**. The received analogue signal *vi* is digitised by an a-to-d converter, and the *p* data bits applied to *p* parallel shift registers. Digital-to-analogue converters at regular intervals along the tapped digital delay line reconstitute delayed versions of *vi*. Four of the outputs are inverted and then the seven signals are summed.

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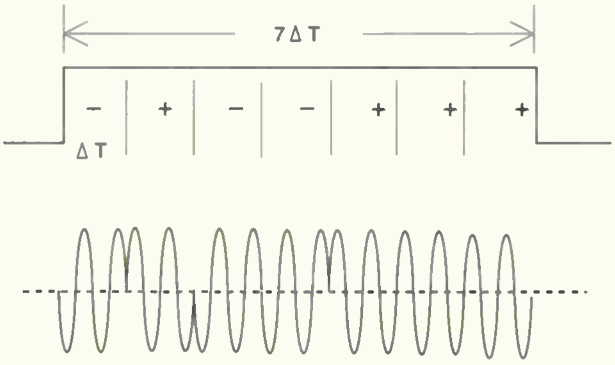


Fig. 1. BPSK modulation with two carrier cycles per 'chip'.

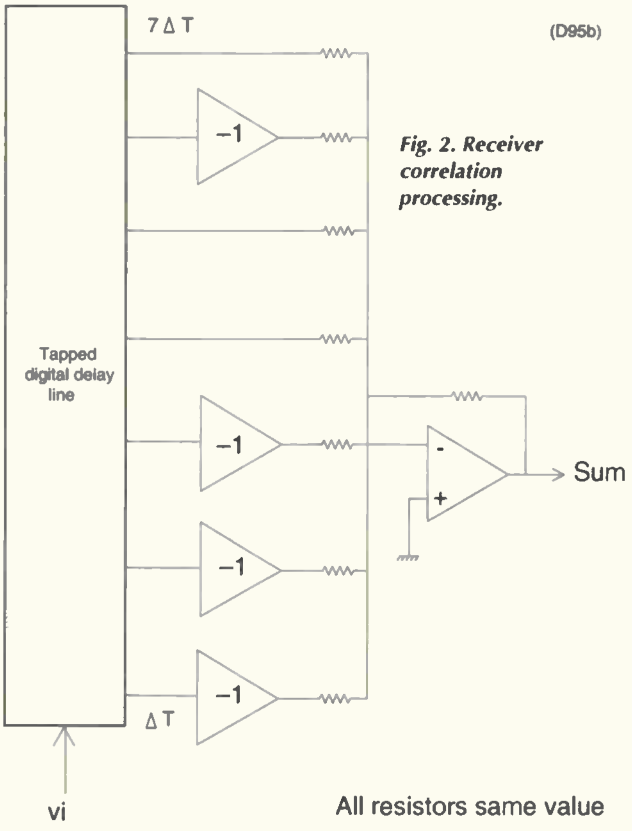
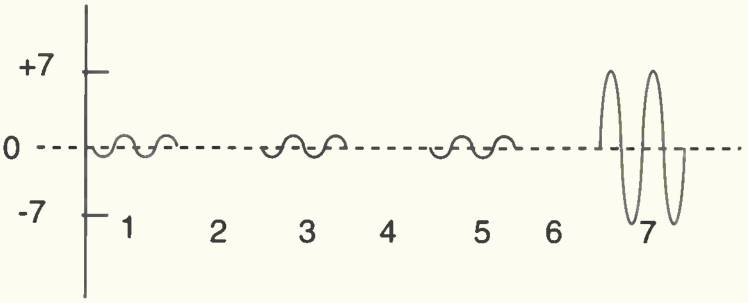


Fig. 2. Receiver correlation processing.

Fig. 3. Correlator output showing wanted signal enhancement.



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National Instruments – computer-based measurement and automation

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Note that the phase code pattern (+++−−+−) in the receiver is the reverse of that used in the transmitter (−+−−+++). The correlation results in an output amplitude of seven times that of the input during the last Δt cell, as in **Fig. 3**.

The scheme can be used either to transmit individual pulses, e.g. radar or sonar responses, etc., or for a bit in a digital transmission, **Fig. 4**.

A parallel-input, serial-output shift register (PISO) is hard wired with a permanent parallel input 0001101 shift code. It has two control inputs, Pi for parallel loading and SR for shifting serially right. A train of TTL clock pulses is derived from the carrier or subcarrier sinewave oscillator, of frequency f_t.

The clock frequency is divided by 2 ($f_c=f_t/2$) and forms the PISO clock input. Frequency f_c is further divided by 7 and ANDed with the serial digital data input Si, to generate Pi for the PISO. Input SR equals /Pi, which is the inverse of Pi.

The output of the PISO operates the controlled inverter; **Fig. 5** shows the timing diagram. The divide by 7 stage may use any of the usual methods, such as a Johnson counter, or JK bistable devices with gating. A sync or reset pulse is used to ensure the pulse Pi occurs at time $7\Delta t$, as in Fig. 5.

A 74166 can be used for the PISO, the unused eighth bit being set to '0'.

K. Balasubramanian
European University of Lefke
Turkish Republic of Northern Cyprus
D95

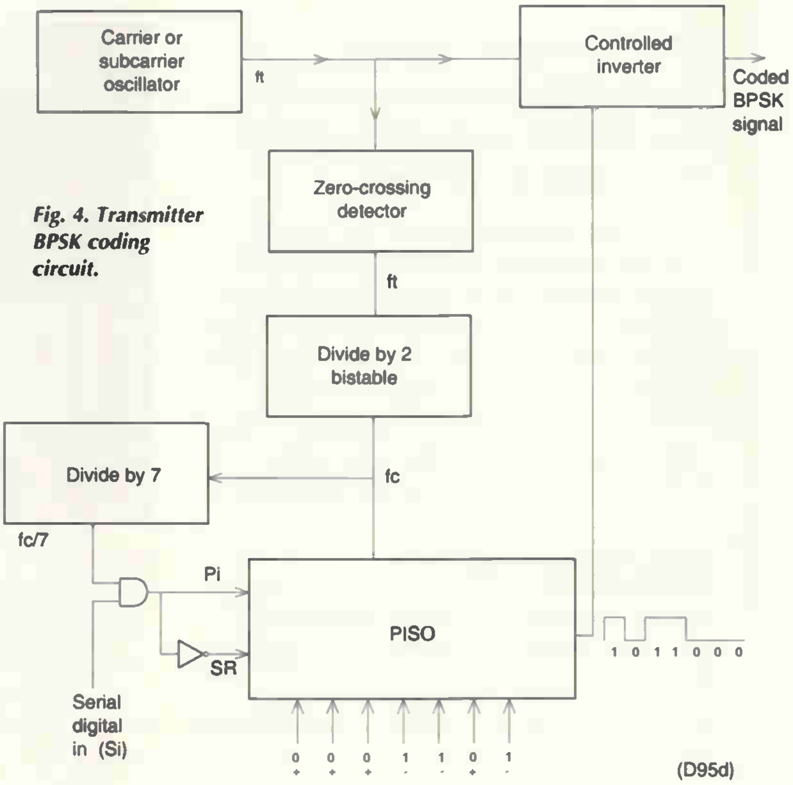


Fig. 4. Transmitter BPSK coding circuit.

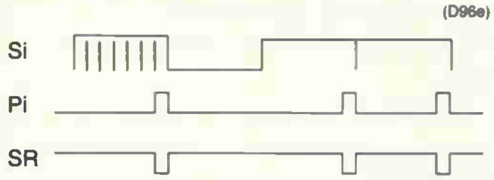


Fig. 5. Timing diagram for Fig. 4.

Comparative light transmission measurements

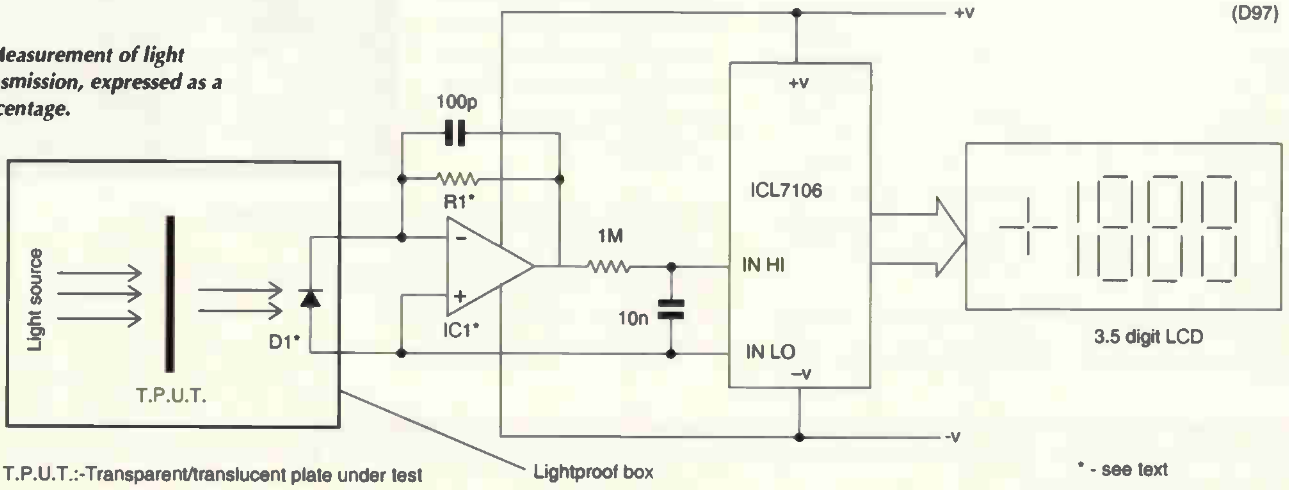
The arrangements described here permit comparative measurements of light transmission, or alternatively of light losses. These losses include reflection and absorption.

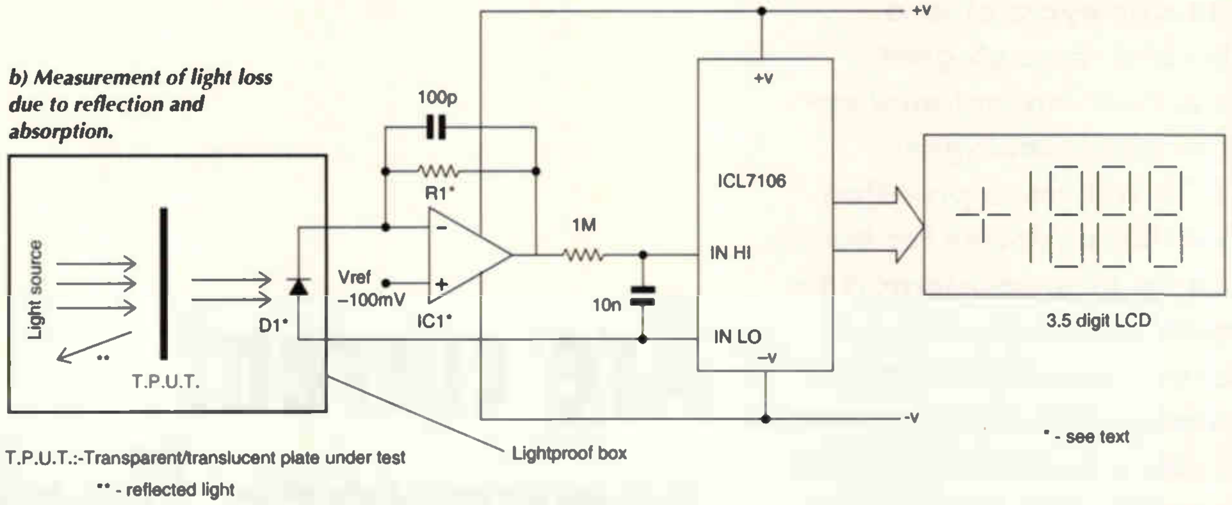
a) The circuit is set up as follows. The ICL7106 reference voltage is set to 100mV. R_1 is chosen such that in the absence of the plate under test, the display reads 100. With the plate under test then in position, the display gives a readout of the percentage transmission through the plate.

b) The arrangement is as in a) except that the non-inverting input of the op-amp is returned to a potential 100mV below 0V (IN LO). Now, the display reads zero in the absence of the plate under test. With a totally opaque plate, the display reads 100 – ideally. Thus the readout gives the light losses (reflected + absorbed light).

Op-amp IC_1 should be a type with negligible input bias current, such as

a) Measurement of light transmission, expressed as a percentage.





b) Measurement of light loss due to reflection and absorption.

T.P.U.T.:-Transparent/translucent plate under test

** - reflected light

* - see text

a CA3130 or MAX406. The interior of the lightproof box should be finished in photographic matt black. The chosen photodiode type should be appropriate to the light source, e.g. for daylight a BPW21 would be suitable. This has a response generally similar to the human eye.

Denis Lenardic
Jesenice
Slovenia
D97

'A' scope target tracking

In radar and sonar applications, the A scope display – received signal vertically versus range horizontally – is popular. The range R is given by $R=c\times\Delta t/2$, where Δt is the time from the transmitted pulse to the reception of the echo, and c is the speed of light, **Fig. 1**. A target tracking system can follow a given target, providing range information continuously. **Figure 2** shows a circuit to perform this function.

Two monostable multivibrators M_2 and M_3 of Fig. 2 produce adjacent time slots, the early gate and late gate, **Fig. 3**. The target echo blip on the CRT display is sampled by these gates, each of width $\tau/2$.

If the areas of the portions of the echo in each gate are equal, then the position of the gates – i.e. the delay Δt relative to the 'tr' pulse, **Fig. 4** – indicates the range. Monostable device M_1 is triggered by the tr pulse, its delay being determined by its timing components. These include an FET, controlled by feedback from the SUM amplifier.

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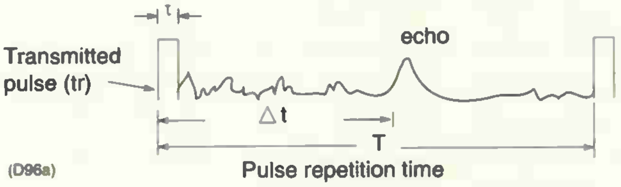


Fig. 1. Echo delay Δt gives the target range.

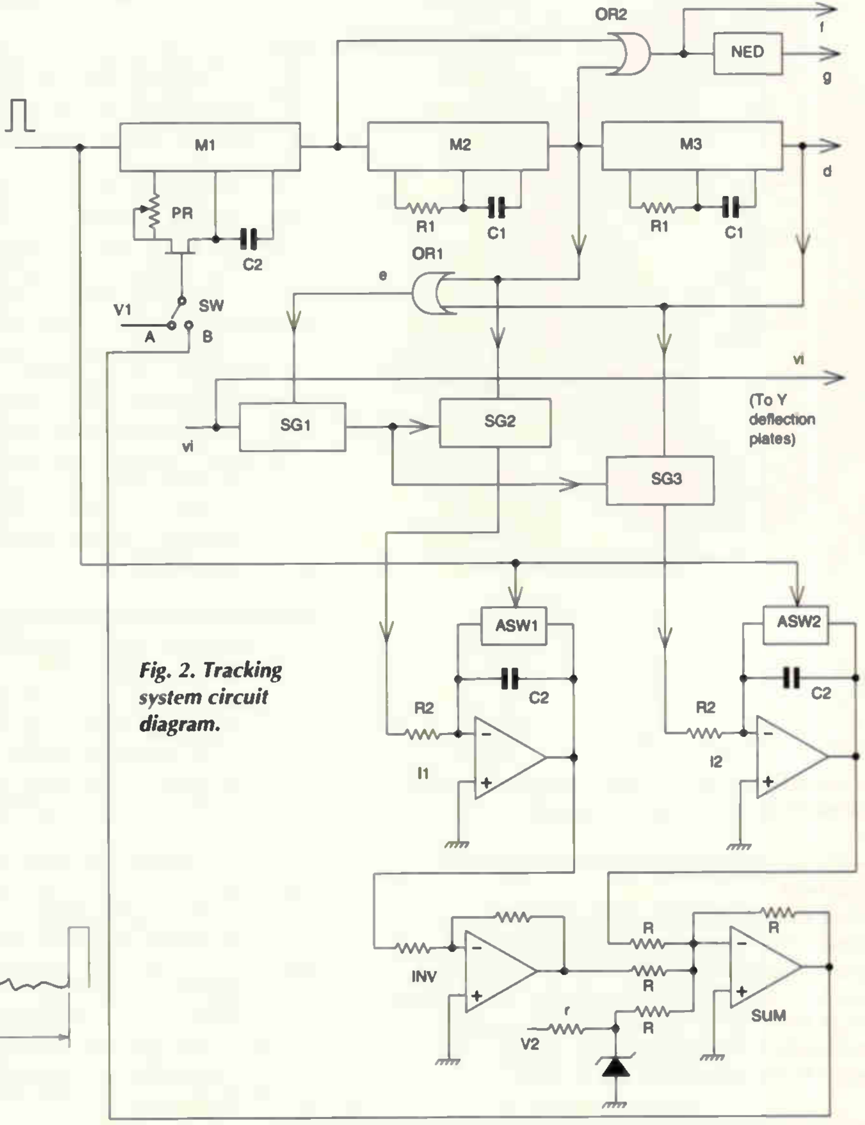
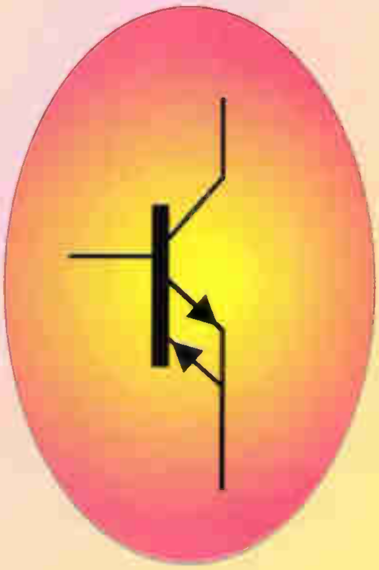


Fig. 2. Tracking system circuit diagram.

Are current conveyors finally coming of age?

Current conveyors offer a simpler and more elegant solution than conventional op-amps in many analogue circuits. A full-wave precision rectifier that performs far better than its op-amp counterpart for example needs only two conveyors, two resistors and four diodes. Monolithic current conveyors have been rare, but recently Burr Brown launched one under the popular guise of 'a perfect transistor'. So now is a good time to get to grips with the circuit concepts.



Not an ideal symbol for the current conveyor, but nevertheless a useful tool for envisaging what it does.

The current-conveyor is an extremely useful general-purpose analogue building block. It is as versatile as the ubiquitous op-amp – maybe even more so.

Until recently, there were no monolithic current-conveyors available. Although Burr Brown does not describe the OPA660 as a current-conveyor, it is just that, with an additional on-chip independent voltage-follower.

In this article the history of the current-conveyor is traced and the inter-relationship between the current-conveyor, the perfect transistor and OTAs is discussed – together with many useful application circuits. A second article looks more closely at the Burr Brown OPA660 – also known as 'The perfect transistor'.

What is a current-conveyor?

The current-conveyor is a general-purpose analogue amplifier. It is a versatile analogue building block. Rather like an ordinary op-amp it can be configured with additional passive components to create a wide range of different analogue signal processing functions.

Where a current-conveyor differs significantly from an op-amp is that it has two internal gain stages – one a unity-gain voltage amplifier, or voltage-follower, the other a unity gain current amplifier, or current-follower. Consequently, as there is no high internal open-loop gain, the current-conveyor is not primarily intended to be used within a negative feedback structure.

The current-conveyor offers an alternative approach to analogue signal processing, leading to new methods of implementing analogue transfer functions. In many cases, conveyor-based designs offer improved performance relative to voltage op-amp-based implementations in terms of accuracy, bandwidth and convenience.

Circuits based on voltage op-amps are generally very straightforward to design since the behaviour of a voltage op-amp can be approximated by a few simple design rules. This is also true for current-conveyors. Once the appropriate design rules are understood, an application engineer is able to design conveyor-based circuits just as easily.

Although the current-conveyor concept has been around for a very long time, it is only relatively recently that general-purpose monolithic current-conveyors are becoming available.

Current conveyor roots

Smith and Sedra proposed the first generation current-conveyor or CCI in 1968[1] and the more versatile second generation current-conveyor, or CCII, was introduced by the same two authors in 1970[2], as an extension of their first generation design.

The CCII is without doubt a much more valuable and adaptable building block and so we will concentrate exclusively on this device. The second generation current-conveyor offers as much, if not more versatility than the operational amplifier.

One particular advantage of the CCII is its current output capability making

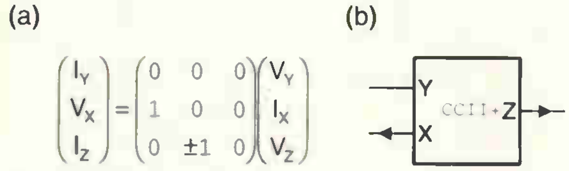

(a)
(b)
$\begin{pmatrix} I_Y \\ V_X \\ I_Z \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 \\ 1 & 0 & 0 \\ 0 & \pm1 & 0 \end{pmatrix} \begin{pmatrix} V_Y \\ I_X \\ V_Z \end{pmatrix}$
Y
X
CCII+
Z

Fig. 1. Second-generation current conveyor, a) describing the I–V matrix and b), the device symbol.

The authors
John Lidgey and Khaled Hayatleh, Oxford Brookes University. The authors are in the Department of Electronic Engineering at Oxford Brookes University, Oxford, UK. John Lidgey is Professor of Electronics and Head, and Dr Khaled Hayatleh is Senior Lecturer in Electronics & Communications. John is also a Co-Director of LTP Electronics Limited, Oxon.

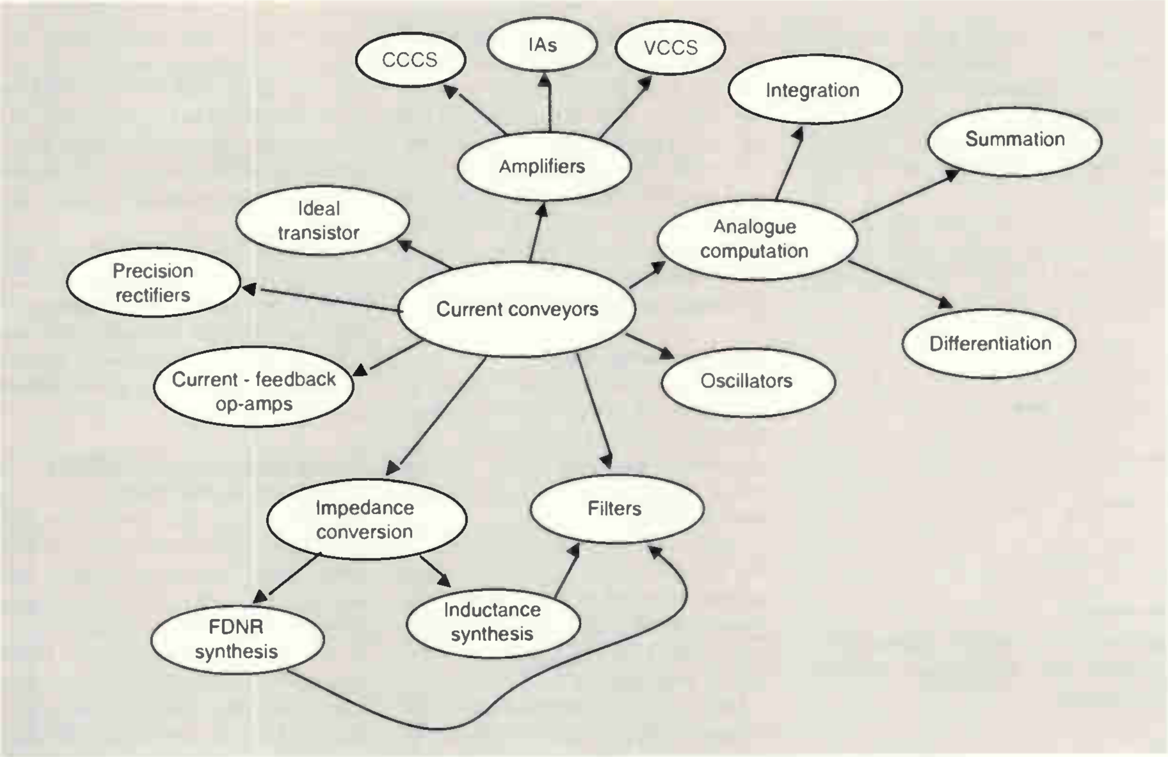


Fig. 2. Current-conveyor application tree.

it ideal for transconductance and current amplifier applications. **Figure 1a)** shows the voltage–current describing matrix for the CCII and Fig. **1b)** the simplified block-schematic representing the CCII. Power supply connections are omitted, as is the normal practice in drawing circuit schematics.

Voltage at the low impedance node *X* follows that at the high impedance input node *Y*. Input current at node *X* is mirrored or 'conveyed' to the high-impedance output node *Z*.

The ± sign indicates the polarity of the output current with respect to the input current. By convention, a positive sign indicates that both the input and output currents simultaneously flow into, or out of, the device. Thus Fig.1b) as shown represents a CCII±.

Note that while nodes Y and Z are clearly a voltage input and current output respectively, node *X* has a dual function – input for current and output for voltage. At first sight this may seem unusual. Because of the combined voltage and current following properties, CCIIs can be used to synthesise a wide-range of analogue circuit functions, some of which are not so easily realisable using conventional voltage op-amps.

In 1968 when the current-conveyor was first thought of it was merely a circuit concept without practical implementation. At that time the semiconductor industry was absorbed with the development of the monolithic voltage op-amp. Without any clearly stated advantages there was little motivation for developing a monolithic current-conveyor.

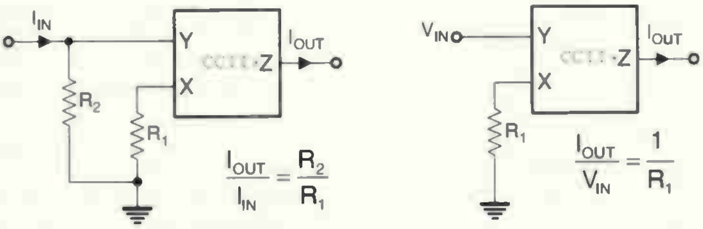

I_{IN}
Y
X
CCII-Z
I_{OUT}
R_2
R_1
$\frac{I_{OUT}}{I_{IN}} = \frac{R_2}{R_1}$
V_{IN}
Y
X
CCII-Z
I_{OUT}
R_1
$\frac{I_{OUT}}{V_{IN}} = \frac{1}{R_1}$

Fig. 3. Current output amplifiers – a) is current amplifier, or more precisely a current-controlled current source, and b) is a voltage-to-current converter or voltage-controlled current source.

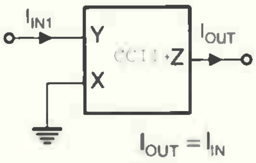


Fig. 4. Current-buffer or current-follower – probably the most useful utility current-mode building block.

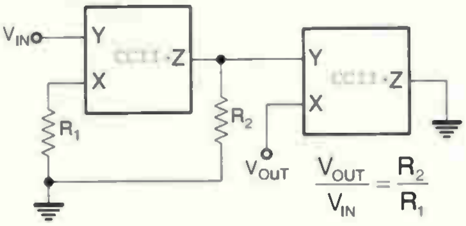


Fig. 5. Voltage amplifier using current conveyors is obtained by adding a voltage-buffer to the voltage-controlled current source of Fig. 3b).

It was not until the development of production, fully complementary bipolar technology, based on dielectric isolation almost 20 years later that the capability of creating a high-performance single-chip CCII became practicable.

Since it was first conceived, several hundred papers have been published on the theory and applications of current-conveyors. Some of these application areas are shown in **Fig. 2**. Next we will take a look at some of these application circuits.

Amplifiers with no overall NFB

The current-conveyor has precise unity current-gain transfer between the *X* and *Z* terminals, rather than the high but ill-defined open-loop gain of the voltage op-amp. As a result, in amplifier applications the current-conveyor is generally used without any overall negative feedback.

The advantage of this approach is that the traditional closed-loop gain-bandwidth conflict of negative-feedback voltage op-amp circuits is avoided. Of course, the benefits of global negative feedback, for example noise reduction, improvement in input and output impedance levels, can no longer be exploited. But the absence of overall negative feedback generally results in wider bandwidths at higher levels of gain with current-conveyor based circuits.

However, to maintain a high level

of accuracy without the use of negative feedback, a high quality current-conveyor integrated circuit realisation is needed with very precise unity gain sections within the device. This is the key design challenge for the IC designer.

The CCII+ can easily be used to create the two current output amplifiers shown in **Fig. 3**. In the current-amplifier, R_2 must be much less than the input impedance at node Y for current transfer accuracy, that is, $R_2 << R_{IN(Y)}$. This factor limits the maximum possible gain of the circuit.

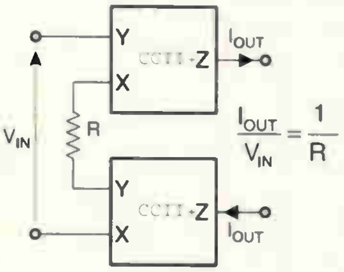


Fig. 6. Differential voltage-to-current converter. Differential input voltage appearing across resistor R generates a current that is conveyed to the two outputs.

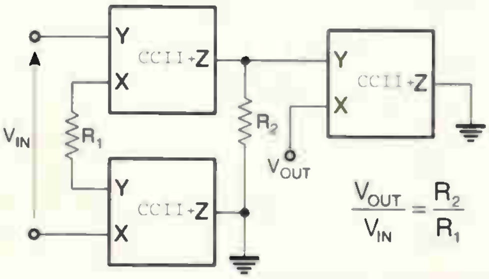


Fig. 7. Current-conveyor instrumentation amplifier has high CMRR, needs no external component matching and its bandwidth is independent of gain.

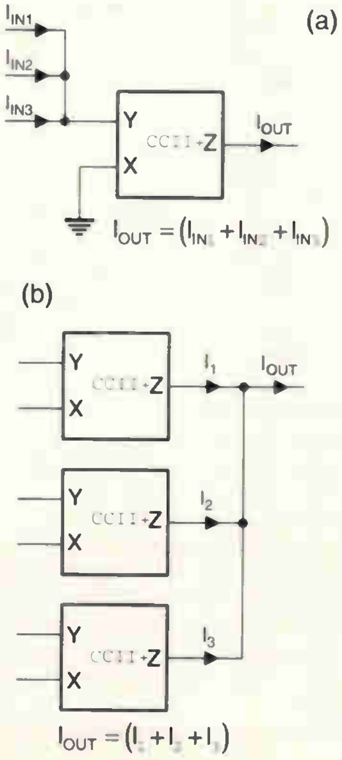


Fig. 8. Current summing with current-conveyors, a) input summing, b) output summing.

In both circuits, R_1 forms a pole with parasitic capacitance to ground at node X. As a result, the value of this resistor should be kept low to ensure that this parasitic pole does not dominate the frequency response.

Current amplifier

The current conveyor can be used to advantage in current-mode circuits. Probably the most useful utility current-mode building block is the unity current gain current amplifier, shown in **Fig. 4**.

There's a low input impedance at the X node, which conveys the input current through to the high-output-impedance Z node. Hence the current gain is unity – negative for a CCII+ and positive for a CCII–.

This circuit is the antithesis of a voltage-follower and is referred to as a current buffer or current-follower.[3] It can be used to advantage as an interface between voltage and current-mode circuits. For example, adding a current-follower before the amplifier of Fig. 3a) provides the circuit with a greatly reduced input impedance. It also improves the integrity of the current drive into the Y node of the second current-conveyor.

Voltage amplification can be obtained by adding a voltage-buffer to the voltage-controlled current source of Fig. 3b). **Figure 5** shows such an amplifier, with a second CCII+ being used as a voltage-follower.

Using two CCIIs, Toumazou and Lidgey [4] described a differential voltage-to-current converter as shown in **Fig. 6**. Any differential input voltage appears across resistor R generates a current in R which is then conveyed to the two outputs.

Provided that both devices are well matched and have a precise voltage-following action between their respective Y and X inputs, the common-mode gain of the circuit will be zero. The circuit does not rely on any external resistor matching for high CMRR. The CMRR will roll-off though at higher frequencies due to parasitic capacitance to ground at the X-nodes, which will then give rise to common-mode currents.

Common-mode performance can be improved by increasing the differential gain that is, by reducing the value of R. The ultimate limit on CMRR will be determined by the mismatch in open-loop bandwidth of the two conveyors.

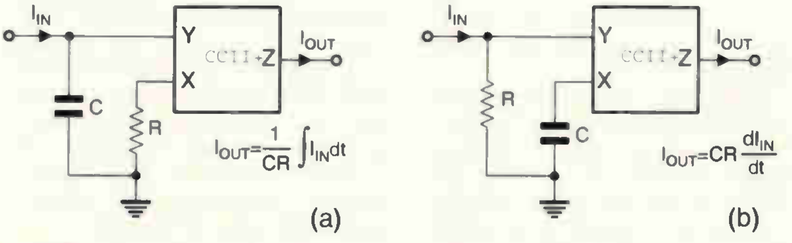
Instrumentation amplifier needs no *R* matching

By converting the output current back to a single-ended voltage, this differential transconductance cell can be extended to produce a high performance instrumentation amplifier[5,6], as shown in **Fig. 7**. This instrumentation amplifier has some highly desirable features – namely a high CMRR, no external component matching requirement and a bandwidth independent of gain.

This instrumentation amplifier has proved to be particularly suitable for the implementation of EMG amplifiers with shutdown control[7].

Shutdown control is necessary to isolate the EMG amplifier during the period that an electrical stimulation is applied to the muscles, since this large stimulus will otherwise saturate the amplifier. The amplifier would then require a recovery period before it is able to measure the much weaker EMG muscle response. This shutdown control is most easily implemented by providing a switch control at the amplifier inputs.

Conventional EMG amplifiers based on voltage op-amps require switching at a high impedance node, and this tends to cause transient spikes which themselves can saturate the amplifier. Using the instrumentation amplifier of Fig. 7, the switching control can be moved to a low impedance node, so that resistor R_1 is either disconnected or connected between the two low impedance X inputs.


I_{IN}
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R
$I_{OUT} = \frac{1}{CR}\int I_{IN}dt$
(a)
I_{IN}
Y
CCII+Z
X
I_{OUT}
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$I_{OUT} = CR\frac{dI_{IN}}{dt}$
(b)

Fig. 9. Current integrators and current differentiators are easily implemented using current conveyors, as illustrated in a) and b) respectively.

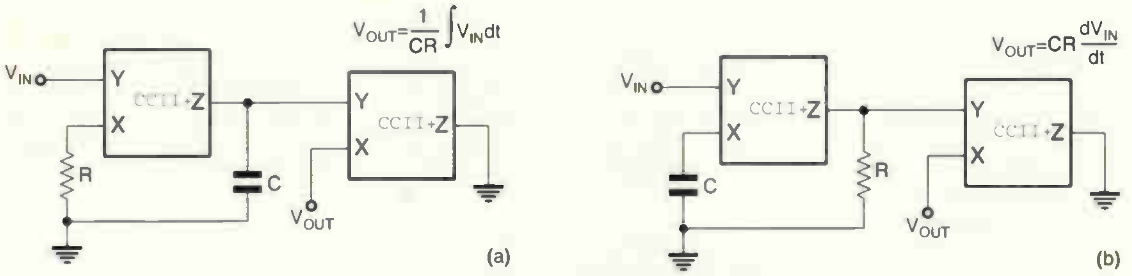
$V_{OUT}=\frac{1}{CR}\int V_{IN}dt$
V_{IN}
Y
X
CCII+Z
R
C
V_{OUT}
(a)
$V_{OUT}=CR\frac{dV_{IN}}{dt}$
(b)

Fig. 10. For voltage integration, a), and differentiation, b), an output voltage buffer is needed to supply load current.

This conveyor-based EMG instrumentation amplifier demonstrates no saturation or recovery problems, and is able to measure the EMG signal accurately, immediately after the stimulation pulse.

Current-mode computation

Current summation is readily achieved with a current-conveyor, either at the low input impedance *X* node or at the high output impedance *Z* node, as shown in **Fig. 8**.

In their original current-conveyor work, Sedra & Smith[1,2] outlined a number of current-mode analogue computational elements. These ranged from integrators and differentiators to more complex function generators.

Current differentiators and integrators are easily implemented, as shown in **Fig. 9**. In Fig. **9a)**, resistor *R* should be fairly small to minimise the effect of stray capacitance at node *X*, meanwhile in Fig. **9b)**, *R* should be limited in size to minimise current transfer errors, that is, $R<<R_{IN(Y)}$.

For voltage differentiation and integration, an additional output voltage buffer is required to supply load current, as shown in **Fig. 10**.

In many cases, circuits containing voltage op-amps with capacitive feedback components must be designed very carefully since a capacitor contributes phase lag, which can result in circuit instability. However, because of the absence of global negative feedback, conveyor-based integrator and differentiator circuits do not suffer the same potential instability problems.

The circuits of Figs 9 and 10 use grounded rather than floating capacitors, which is an advantage for integrated circuit realisations.

Impedance conversion

Many impedance converters based on voltage op-amps need complicated circuitry and tight component matching. With its combined voltage and current following properties, the CCII is a more flexible alternative to implement these functions.

Impedance converters are used in many areas of active circuit design, including active filters and inductance simulation. The frequency-dependent negative resistor, or FDNR, is a circuit element used in higher-order filter design that is synthesised using impedance converters.

Components with negative impedance can be used to cancel positive lossy elements in certain application circuits by generating an identical negative component.

Figure 11a) shows a negative impedance converter presented by Sedra & Smith.[2] By terminating input node *Y* with an impedance Z_T as in Fig. **11b)**, a current-controlled NIC, or negative admittance converter is obtained. Similarly, by terminating node *X* with an impedance of *ZT*, as shown in Fig. **11c)**, and driving input node *Y* with a voltage, a voltage controlled NIC is obtained.

Current-conveyor absolute value circuits

Another elegant use of the current-conveyor is for high-speed precision rectification or absolute-value circuit design.

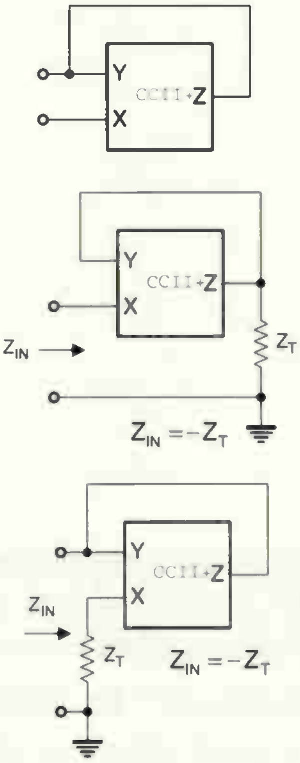

Fig. 11. Grounded negative-impedance converters. Sedra & Smith first presented circuit a). By terminating input node Y with an impedance Z_T as in b), a current-controlled negative-admittance converter is obtained. Similarly, by terminating node X with an impedance of Z_T as in c) and driving input node Y with a voltage, a voltage-controlled NIC is obtained.

The classical problem with conventional precision rectifiers based on diodes and op-amps is that during the non-conduction/conduction transition of the diodes the op-amps have to recover with a finite small-signal *dV/dt*. This results in significant distortion during the zero-crossing of the input signal.

Using high slew-rate op-amps does not solve this fundamental drawback since it is a small-signal transient problem. Conventional rectifiers are thus limited to a frequency performance well below the gain-bandwidth product of the amplifier.

A full-wave precision rectifier can

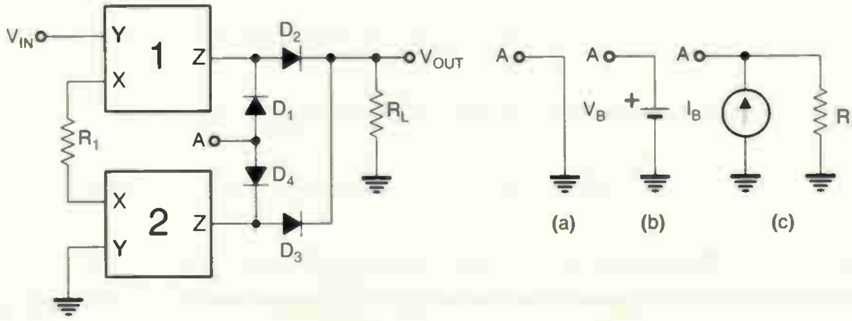

Fig. 12. Precision full-wave rectifier built using current conveyors needs few components yet performs well at high frequencies relative to its voltage op-amp counterpart. Biasing a) is PFWR proposed in ref. 8, b) is voltage biasing technique proposed in ref. 9 and c) is an alternative current biasing technique, ref. 10.

be configured easily using two CCIIs, as in **Fig. 12**. Both the CCIIs form a differential voltage-to-current converter such that during the positive input cycle, the output currents of value V_{IN}/R flows out of the Z node of CCII (1) and into the Z node of CCII (2). As a result, only D_4 and D_2 are active.

Unity gain is achieved if $R_1=R_L=R$. Because D_2 is active, current from the Z node of CCII (1) flows into the output resistor R_L making $V_{OUT}=V_{IN}$. During the negative input cycle, only D_3 and D_1 are active so the output current of CCII (2) is driven into R_L, making $V_{OUT}=-V_{IN}$. Clearly the magnitude of gain is R_L/R_1 and this can be increased from unity by making R_L greater than R_1.

Compared with classical op-amp based designs the performance is good. You would naturally expect the circuit to work close to the f_T of the device. But this is not the case. At very low signal levels all the diodes are off. As a result the differential voltage to current converter is transformed into a high gain differential voltage amplifier.

The solution to this problem is to modify the circuit by offsetting the output of the conveyors by biasing all the diodes with bias circuits (b) or (c). ■

John and colleagues investigate the Burr Brown OPA660 'Perfect transistor' in their next article.

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Ensuring oscillator signal quality

A multitude of tips for enhancing purity and stability of RF oscillators, presented by Joe Carr.

An ideal sine wave RF oscillator ideally produces a nice, clean harmonic and noise-free output on a frequency that is stable. But real RF oscillators tend to have certain problems that detract from the quality of their output signals.

Load impedance variation and DC power supply variation can both cause the oscillator frequency to shift. Temperature changes also cause frequency change problems. Noise that modulates the oscillator, both externally and internally generated, produces phase noise sidebands around the oscillator signal.

An oscillator that changes frequency without any help from the operator is said to drift. Frequency stability refers generally to freedom from frequency changes over a relatively short period of time – for example a few seconds to dozens of minutes. This problem is different from ageing, which refers to frequency change over relatively long periods of time – i.e. hertz/year – caused by ageing of the components.

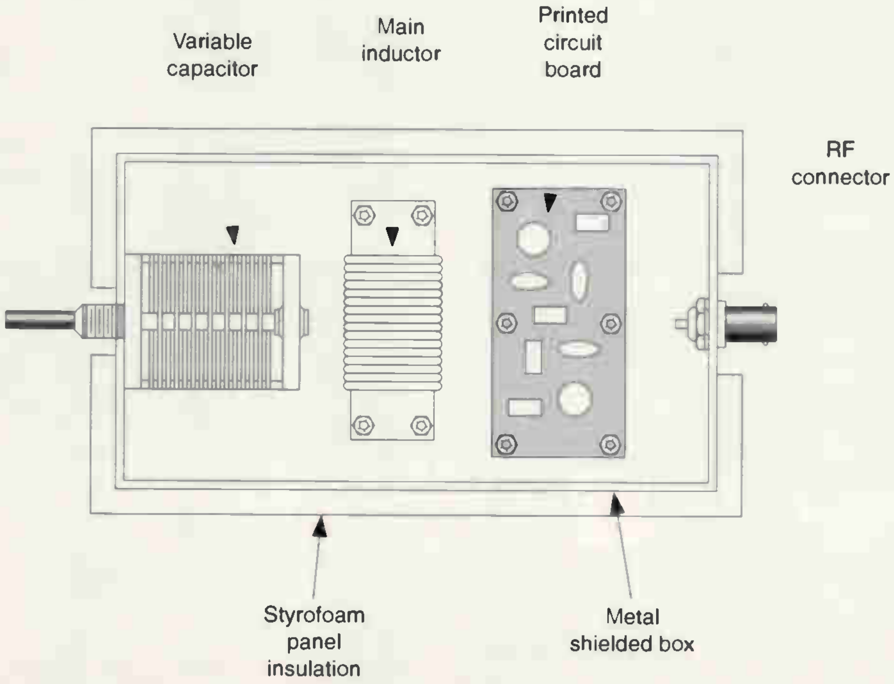


Fig. 1. Thermally isolating variable-frequency oscillator circuitry reduces short-term drift caused by changes in ambient temperature.

Temperature changes are a large contributor to this problem in two forms. These are warm-up drift – i.e. drift in first 15 minutes – and ambient temperature change drift.

If certain guidelines are followed, then it is possible to build a very stable oscillator. For the most part, the comments below apply to both crystal oscillators and *LC* tuned oscillators, although in some cases one or the other is indicated by the text.

Temperature effects

Temperature variation has a tremendous effect on oscillator stability. Avoid locating the oscillator circuit near any source of heat within the equipment where it serves. In other words, keep it away from power transistors or IC devices, voltage regulators, rectifiers, lamps or other sources of heat.

Thermal isolation. One approach is to thermally isolate the oscillator circuits. **Figure 1** shows one such method. The oscillator is built inside a metal shielded cabinet as usual, but has styrofoam insulation applied to the sides.

There is a type of poster board which has a backing of styrofoam to give it substance enough for self-support. It is easily cut using a scalpel or the like.

Cut the pieces to size, and then glue them to the metal surface of the shielded cabinet using contact cement, or some other form of cement that will cause metal and paper to adhere together.

Crystal ovens. When very good temperature stability is required, it is common practice to use a constant-temperature oven for crystal oscillators, **Fig. 2**. The oven keeps the crystal at a constant temperature of 75°C or 80°C.

There are two basic forms: snap action and proportional. The snap action oven uses a thermal switch that turns on and off at pre-set temperatures, much like the thermostat in a centrally heated house. The proportional oven uses a temperature sensor and circuitry to provide heating as needed to keep the temperature stable.

A number of integrated circuits are produced that serve as proportional controllers. Both socket mounted and printed circuit ovens are available – often available from crystal manufacturers.

Self heating. Operate the oscillator at as low a power level as can be tolerated so as to prevent self-heating of the active device and associated frequency determining components. It is generally agreed that a power level in the order of 1 to 10mW is sufficient.

If a higher power is needed, then a buffer amplifier can be used. The buffer amplifier also serves to isolate the oscillator from load variations.

Important criteria

Use low frequencies. In general, *LC* controlled variable-frequency oscillators should not be operated at frequencies above about 12MHz. For higher frequencies, it is better to use a lower frequency variable-frequency oscillator and heterodyne it against a crystal oscillator to produce the higher frequency.

For example, one common combination uses a 5 to 5.5MHz main variable-frequency oscillator for all HF bands in SSB transceivers.

Feedback level. Use only as much feedback in the oscillator as is needed to ensure that the oscillator starts quickly when turned on, stays operational, and does not 'pull' in frequency when the load impedance changes.

In some cases, there is a small value capacitor between the *LC* resonant circuit and the gate or base of the active device. The reason for using a small value capacitor is to prevent drift by lightly loading the tuned circuit.

The most common component for this job is a 3 to 12pF NP0 disc ceramic capacitor. Adjust its value for the minimum level that ensures good starting and freedom from frequency changes under varying load conditions.

Output isolation. A buffer amplifier – even if it is a unity gain emitter follower – is highly recommended. It will permit building up the oscillator signal power, if that is needed, without loading the oscillator. The principle use of the buffer is to isolate the oscillator from variations in the output load conditions.

Figures 3 and 4 show typical buffer amplifier circuits. The circuit in Fig. 3 is a standard emitter follower – i.e. common collector – circuit. It produces near-unity voltage gain, but some power gain.

Another type of buffer amplifier is shown in Fig. 4. It is a feedback amplifier using a 4:1 balun-style toroidal core transformer in the collector circuit. This circuit exhibits an input impedance near 50Ω, so should only be used if the oscillator circuit can drive 50Ω. In some cases, a higher impedance circuit is needed.

DC power supply. Power supply voltage variations have a tendency to frequency modulate the oscillator signal. Because dynamic circuit conditions often result in a momentary transient droop in the supply voltage, and because line voltage variations can cause both transient drops and peaks, it is a good idea to use a voltage regulated DC power supply on the oscillator.

It is a very good idea to use a voltage regulator to serve the oscillator alone – even if another voltage regulator is used to regulate the voltage applied to other circuits, **Fig. 5**. Although this double regulation approach may be a cost burden, it is reasonable, given the cost of three-terminal IC voltage regulators and the advantage gained.

For most low-powered oscillators, a simple low-power 'L-series' – for example a 78L08 – three-terminal integrated circuit voltage regulator for IC_2 in Fig. 5 is sufficient. The L-series devices provide up to 100mA at the specified voltage, which is enough for most oscillator circuits.

Capacitors on both input and output sides of the voltage regulator – C_1 and C_2 in Fig. 5 – add further protection from noise and transients. The values of these capacitors are selected according to the amount of current drawn.

The idea is to have a local supply of stored current to temporarily handle sudden demand changes, allowing time for

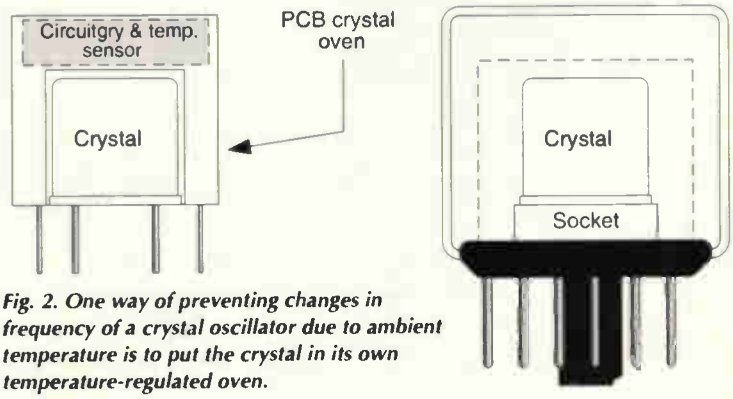


Fig. 2. One way of preventing changes in frequency of a crystal oscillator due to ambient temperature is to put the crystal in its own temperature-regulated oven.

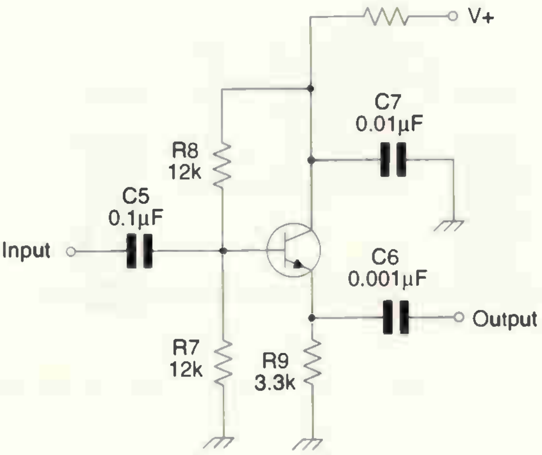


Fig. 3. Loading affects the performance of an oscillator and adding an emitter follower buffer like this is highly recommended.

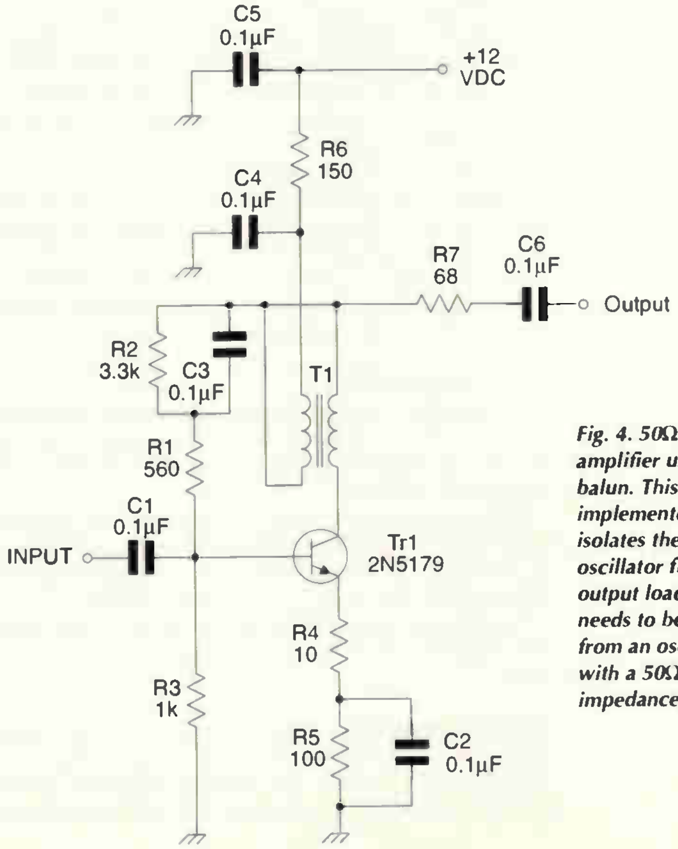


Fig. 4. 50Ω buffer amplifier using a 4:1 balun. This easily implemented circuit isolates the oscillator from the output load, but it needs to be driven from an oscillator with a 50Ω output impedance.

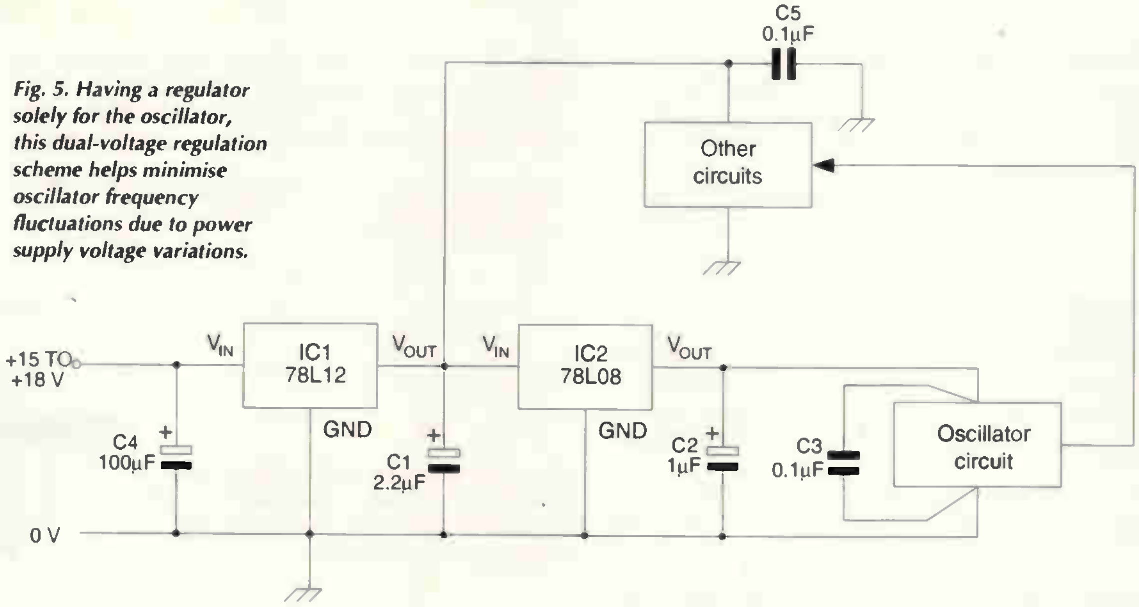


Fig. 5. Having a regulator solely for the oscillator, this dual-voltage regulation scheme helps minimise oscillator frequency fluctuations due to power supply voltage variations.

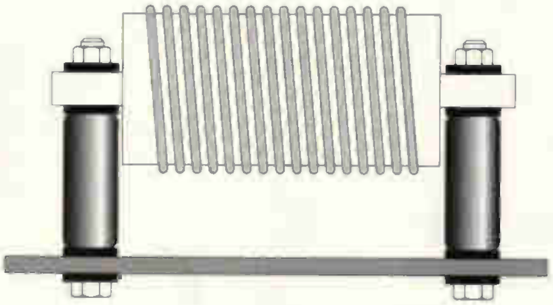


Fig. 6. To reduce the vibration sensitivity of an LC oscillator, the inductor should be fixed to the PCB via sock absorbers.

a transient to pass, or the regulator to 'catch-up' with the changed situation.

Vibration isolation. Frequency determining components of the oscillator can affect its stability performance.

The inductor should be mounted so as to prevent vibration. While this requirement means different things for different styles of coil, it is nonetheless important. Some people prefer to mount coils rigidly, while others will put them on a vibration-isolating shock absorber material.

I've seen cases where a receiver that required a very low SSB noise sideband local oscillator was not able to meet specifications until some rubber shock absorbing material was placed underneath the oscillator printed wiring board.

Any time that vibration affects crystals, inductors or capacitors, there is a possibility of high vibration-induced noise on the oscillator output signal.

Coil considerations

Core selection. Air-cored coils are generally considered superior to those with either ferrite or powdered iron cores because the properties of magnetic cores are affected by temperature variation.

Of those coils that do use cores, slug-tuned are said to be best because they can be operated with only a small amount of the tuning core actually inside the windings of the coil, reducing the vulnerability to temperature effects.

Toroidal cores have a certain endearing charm, and can be used wherever the ambient temperature is relatively constant.

Coil-core processing. One source recommends tightly winding the coil wire onto the toroidal core, and then heat annealing the assembly. This means placing it in boiling water for several minutes, and then removing and allowing it to cool in ambient room air while sitting on an insulated pad. I haven't personally tried it, but the source did and reported remarkable freedom from inductor-caused thermal drift.

For most applications – especially where the temperature is relatively stable – the coil with a magnetic core can be wound from enamelled wire. Usually, #20 to #32 AWG are specified. For best stability though, it is recommended that Litz wire be used.

Although Litz wire is a bit hard to obtain in small quantities, it offers superior performance over relatively wide changes in temperature. Be aware that this nickel-based wire is difficult to solder properly, so be prepared for a bit of frustration.

Air-cored coils. For air-cored coils, use #18 SWG or larger bare solid wire, wound on low temperature coefficient of expansion formers. **Figure 6** shows how the coil can be mounted. Stand-off insulators provided adequate clearance for the coil, and hold it to the chassis. The mount shown in Fig. 6 has shock absorbers to prevent movement.

Capacitor Selection

Trimmer capacitors used in the oscillator circuit should be air dielectric types, rather than ceramic or mica dielectric trimmers, because of their lower temperature coefficient.

The small fixed capacitors used in the oscillator should be NP0 disc ceramics (i.e. zero temperature coefficient), silvered mica or polystyrene types.

Some people dislike the silvered mica types because they tend to be a bit quirky with respect to temperature coefficient. Even out of the same batch they can have widely differing temperature coefficients on either side of zero.

Sometimes, you will find fixed capacitors with other than zero temperature coefficient in an oscillator frequency determining circuit. These are used to make temperature-compensated oscillators. The temperature coefficients of certain critical capacitors are selected to create a counter drift that cancels out the natural drift of the circuit.

The main air-dielectric variable capacitor used for tuning should be an old-fashioned double bearing type, i.e. it should have a bearing on each end-plate. For best results, it should be made with either brass or iron stator and rotor plates – not aluminum – and it should be ruggedly constructed.

Compensating for temperature coefficient

Temperature-compensated crystal oscillators can be built by using the temperature coefficient of some of the capacitors

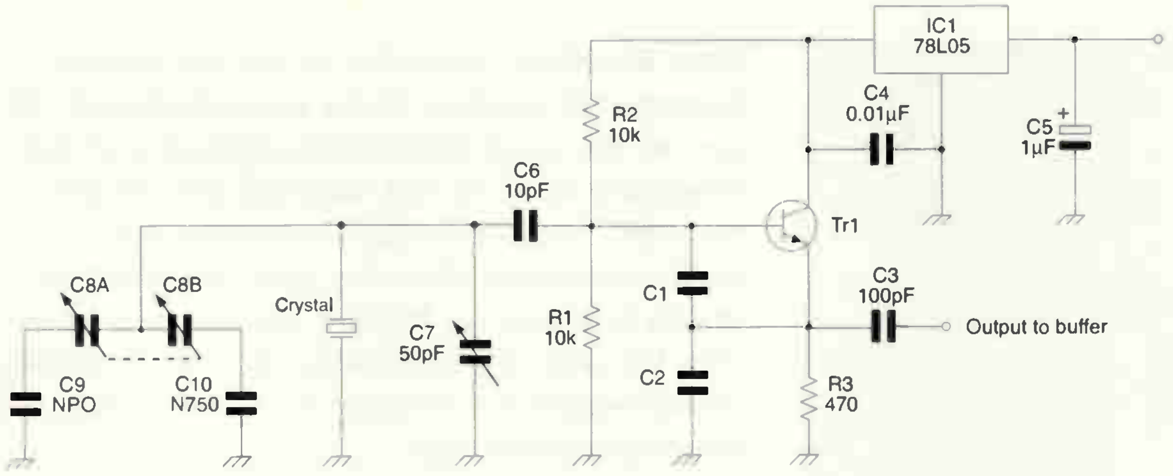


Fig. 7. By careful choice of capacitor temperature coefficients, oscillator frequency drift due to the temperature coefficient of the crystal can be nulled.

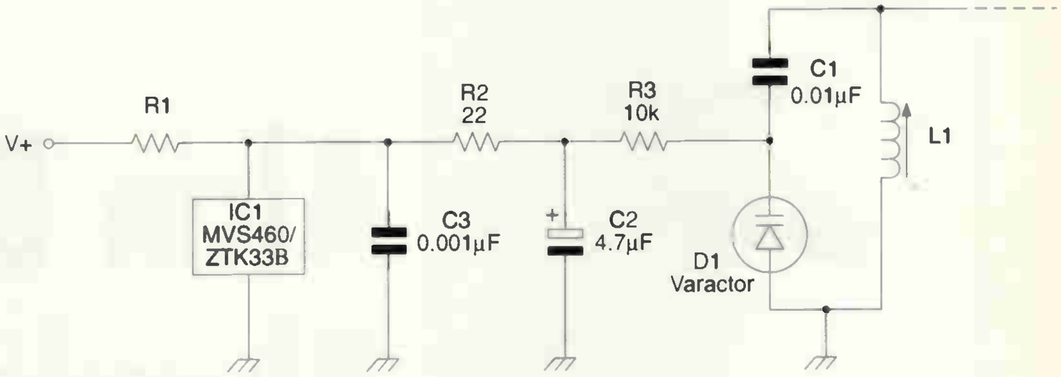


Fig. 8. Arrangement for compensating temperature drift in the varactor involves an MVS460 which has a –2.3mV/°C temperature coefficient.

in the circuit to cancel drift in the opposite direction.

The circuit in **Fig. 7** was used in an amateur radio transceiver at one time. It is a standard Colpitts crystal oscillator, but with the addition of a temperature coefficient cancellation circuit (C_{8A}, C_{8B}, C_9 and C_{10}).

The key to this circuit is C_8, which is a differential variable capacitor, i.e. it contains two sections connected reciprocally so that one is increasing capacitance as the other is decreasing capacitance as the shaft turns.

In most cases, C_9 and C_{10} have the same capacitance, but one is NPO and the other N750 or N1500. If C_8 is differential, and C_9=C_{10}, then the net capacitance across the crystal is constant with changes in C_8 shaft rotation. But the net temperature coefficient does change. The idea is to crank in the amount of temperature coefficient that exactly cancels the drift in the circuit's other components.

The problem with this circuit is that differential capacitors are quite expensive. It may be useful, however, to connect the circuit as shown, and find the correct setting. Once the setting is determined, disconnect $C_{8\text{-}10}$ without losing the setting, and measure the capacitance of the two sections. With this information you can calculate the capacitances required for each temperature coefficient and replace the network with appropriately selected fixed capacitors.

In general, when selecting temperature compensating capacitance, keep in mind the following relationship for the temperature coefficient of frequency:

$$TC_f = \frac{-\left[TC_l + \frac{TC_{C1} \times C_1}{C_{Total}} + \frac{TC_{C2} \times C_1}{C_{Total}}\right]}{2}$$

Where TC_f is the temperature coefficient of frequency, TC_L is the temperature coefficient of the inductor, TC_{C1} is the temperature coefficient of C_1, TC_{C2} is the temperature coefficient of C_2 and C_{Total} is the sum of all capacitances in parallel with L_1, including C_1, C_2 and all strays and other capacitors. This equation assumes that two capacitors, C_1 and C_2, are shunted across an inductor in a variable-frequency oscillator circuit.

Varactors

Voltage-variable capacitance diodes, or 'varactors', are often used today as a replacement for the main tuning capacitor. If this is done, then it becomes critical to temperature control the environment of the oscillator.

It seems that temperature variations will result in changes in diode p-n junction capacitance, and that contributes much to thermal drift. Varactor temperature coefficients of 450ppm are not unusual.

A good way to stabilise these circuits is to supply a voltage regulator that has a temperature coefficient that is opposite to the direction of the varactor drift. By matching these two, it is possible to cancel the drift of the varactor.

Figure 8 shows a basic circuit that will accomplish this job. Diode D_1, in series with DC blocking capacitor C_1, is in parallel with the inductor L_1. These components are connected into an oscillator circuit, which is not shown for sake of simplicity.

Normally, resistor R_3, which is typically 10kΩ to 100kΩ, is used to isolate the diode from the tuning voltage source. It is connected to the wiper of a potentiometer that varies the voltage.

In this circuit though, an MVS-460 (USA type number) or ZTK33B (European type number) varicap voltage stabiliser is present. This device is similar to a zener diode, but has the required –2.3 mV/°C temperature coefficient. It operates at a current of 5mA at a supply voltage of 34 to over 40V DC. The value of resistor R_1 is found from.

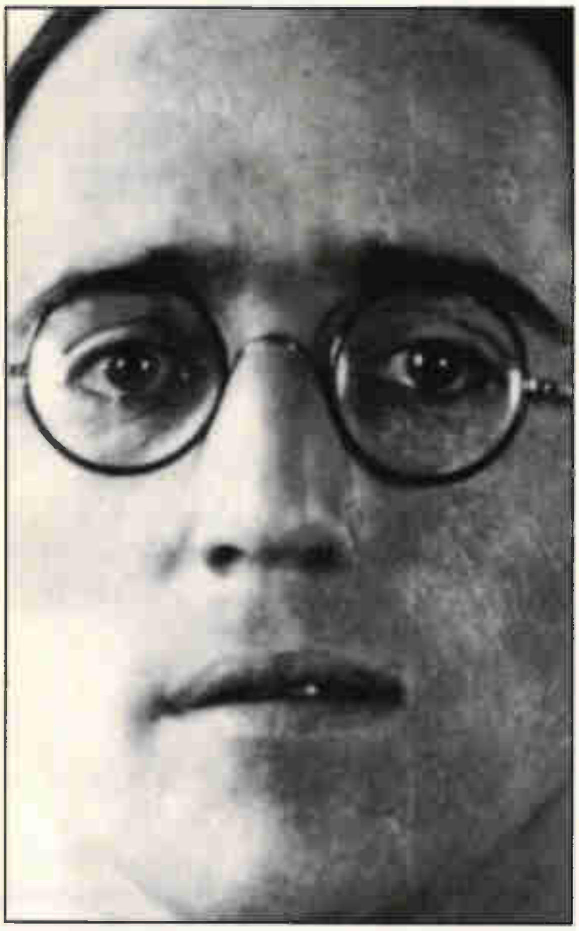
$$R_1 = \frac{(V+) - 33\text{V}}{0.005\text{A}}$$

For example, when V+ is 40 volts, then the resistor should be 1400Ω, or in practice, 1.5kΩ.

In summary

Oscillator circuits should be stable for best operation. If you follow these guidelines, then it is likely that your circuits will be highly successful.

While these techniques do not exhaust the possibilities for stable oscillator construction, they are a good start. These represent a practical collection of weapons for your use. ■



Alan Blumlein – inventor of stereo and co-inventor of modern radar and television – is one of the most significant engineers of the twentieth century. Yet following his death, his work was shrouded in secrecy. He received neither obituary nor tributes. This article is based on Robert Alexander's book 'The life and works of Alan Dower Blumlein', which represents the first comprehensive Blumlein biography.

Alan Dower Blumlein

Shortly after 4.20 in the afternoon on Sunday, 7 June 1942 – a glorious summer's day, clear skies, warm sunshine and perfect visibility for flying – a Halifax bomber crashed into the steep hillside of a valley just north of the River Wye near the village of Welsh Bicknor in Herefordshire. All of its eleven occupants were killed in the enormous fire that engulfed the aircraft on impact.

Of the scientific personnel who died that day, Alan Dower Blumlein stands out as possibly the greatest loss. "A national tragedy," one of his colleagues would call it. At a time when scientific genius was at its foremost, Blumlein was, without a doubt, one of the most brilliant engineers of the twentieth century.

Chain Home

Chain Home was a series of 300-foot high radio transmission and reception towers which started to appear at strategic points along the coastline of Britain from late 1937. Eventually they stretched from Scotland in the North right around the coast as far as Cornwall.

Constructed during the last few years of peace, the Chain Home system was finished just in time for the outbreak of war in September 1939. It played a vital role during the Battle of Britain the following summer.

The system gave enough of an early warning for the RAF Spitfires and Hurricanes to intercept with great accuracy and speed the attacking German aircraft as they approached.

Born in Hampstead in June 1903, Blumlein had graduated from City & Guilds in 1921 with a first-class degree in heavy electrical engineering. This in itself would not bear mention were it not for the fact that by the age of thirteen, the precocious and often eccentric young Blumlein could still not read and write.

He simply found no need to be able to write. As with all things in his life up to this time, if he saw no need, he showed no interest. It was only through sheer determination upon realising that in order to advance his passion for everything electrical, that Alan Blumlein set himself the task of learning to read detailed reference books on his chosen subject.

After a slow start...

Blumlein's career initially took gradual steps. In 1925, he co-published a somewhat elementary paper on basic electrical principles in *Wireless World*.

Though he presented his work the following year to the IEE and was subsequently awarded for it, Blumlein would only return once to the printed word in order to enlighten the world of his thinking.

Following a short but eventful career with Standard Telephones & Cables, during which he obtained the first of his 128 patents, Alan Blumlein applied for a position at The Columbia Graphophone Company. This was in early 1929. While at STC he would meet his employer, mentor and later friend, Isaac Shoenberg, who later became Sir Isaac.

Shoenberg was looking for an engineer to design and construct a recording mechanism that could overcome the patent that Bell Laboratories was imposing on everybody in the record making business.

Blumlein set about designing the elements of a recording and reproducing system. By 1930, this system had successfully bypassed Bell's and would go on to earn Columbia a fortune.

One day in 1931, while at the cinema with his fiancée Doreen, Blumlein enquired of her if she had noticed how the voice for the person on the screen only ever came from one place. Not being of a technical nature, Doreen said that she had not. "Well, I have a way of making the voice follow the person", Blumlein replied.

This casual remark was the first indication of the train of thought which would lead to Alan Blumlein's 'Binaural Sound' patent – arguably his best – and certainly to become one the most important advances in audio engineering of the twentieth century.

But humans have two ears...

Binaural Sound is of course known today as stereo. It works on the basis that human beings have two ears which, because of their orientation on the head, receive sound at slightly different times.

Alan Blumlein ingeniously accommodated the basic concept of binaural sound using electronic circuitry and two loudspeakers. Unfortunately, it was so far ahead of its time in 1931, that many of his colleagues at EMI, did not realise its full potential. EMI had been formed earlier the same year when Columbia and HMV had merged by the way.

Blumlein continued with this work for several more years. He made the first stereo recordings and also the first stereo films before binaural was shelved for a more enlightened time.

Quest to develop TV

By this time, EMI had become involved in the quest to develop a feasible television service.

In 1934, the government formed a committee to investigate the potential of television. This committee concluded that a British television service should be developed by the end of 1936.

Two companies stood out among those tendering systems for a television service. Baird Television, founded by John Logie Baird, was one of them. He had persisted with a mechanical projection method. Despite its ingenious complexity though, this system produced poor quality pictures.

The other company, Marconi-EMI, had decided to work with an all-electronic method of picture transmission and reception. It involved cathode-ray-tube technology, which was then still in its infancy.

Several seemingly insurmountable problems presented themselves to these pioneers. Not least of these was that in many cases the entire electrical circuitry of the system needed to be invented from scratch.

Luckily, EMI possessed an extraordinary set of individuals who, as an engineering team, managed to invent, construct and demonstrate a fully working television system in the now quite unbelievable period of just fourteen months.

As leader of the team in charge of developing the circuitry for the new system, Alan Blumlein had possibly the most enormous task. Yet from this period of his life more than half of his 128 patents were to emerge with many of them critical to the eventual 405-line television system that the BBC adopted.

November 1936 saw the start of a three month trial involving transmissions from Alexandra Palace. The Baird and Marconi-EMI systems transmitted on an alternate basis.

By spring 1937, following the conclusion of the trial, the government and the BBC chose the Marconi-EMI system as it had proved far superior to Baird's.

Who invented television?

It is a curious irony that to this day many consider John Logie Baird to be the inventor of the television. Baird himself never actually claimed this. His mechanical television system proved inadequate for transmission.

It was the team at EMI, whose numbers included Alan Blumlein, that should be given the credit for the 'invention' of the system that we know as television.

As a testimony to the team's work, the 405-line transmissions actually continued until 1986, much as they had during those first trials at Alexandra Palace some fifty years earlier. Originally, the 405-line service was only intended to run for a few years before being updated.

With war in Europe looming, much attention was being directed towards a method of early warning against attack from the air.

The first practical method of electronic radio detection finding – or Radar as it would eventually be known – had been demonstrated by Robert Watson Watt in early 1935. These experimental radio detection finding systems were shrouded in enormous secrecy. They, and their subsequent developments, had led to the construction of the Chain Home system, described in the panel.

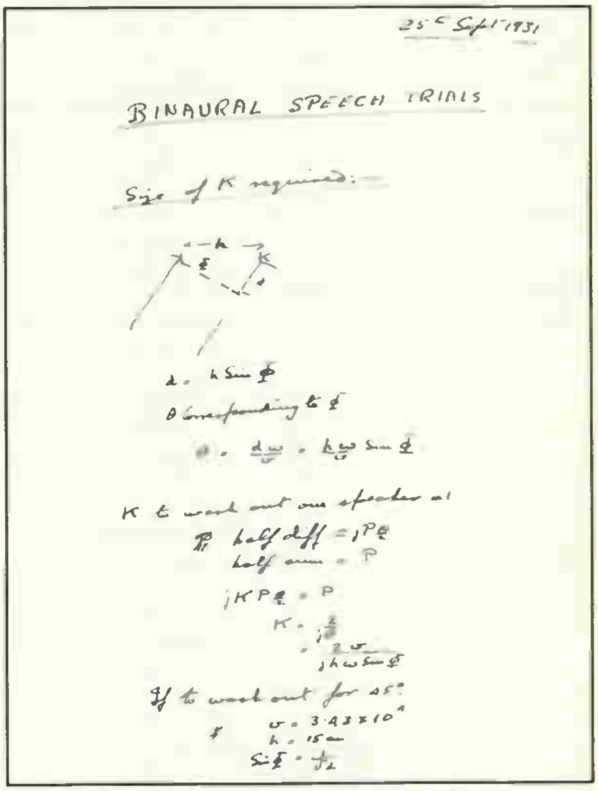
Sonic 'radar'

There were methods other than radio that could be used for detecting approaching aircraft though. It was these alternative solutions that EMI and Alan Blumlein now found themselves interested in.

Sound-detection systems had been tried during the First World War. By 1938, larger, more effective 'ears' to the sky were being experimented with.

Blumlein realised that by incorporating the basis of his binaural sound system in to an aircraft sound detector, a much more accurate fix could be obtained. If this were then displayed on

Unveiled in 1977, this Blue Plaque at 37 The Ridings, Ealing, just off Hanger Lane, was the first – and currently only – permanent memorial to Alan Dower Blumlein that has been erected since his premature death. The Blumleins moved here in 1938, and Alan Blumlein lived and worked at the house until his death in 1942.



25th Sept 1931

BINAURAL SPEECH TRIALS

Size of K required:—

$d = h \sin \Phi$

θ corresponding to Φ

$\theta = \frac{d\omega}{v} = \frac{h\omega}{v} \sin \Phi$

K to wash out one speaker at

Φ half diff $= jP\theta$

half sum $= P$

$jKP\theta = P$

$K = \frac{1}{j\theta}$

$= \frac{v}{jh\omega \sin \Phi}$

If to wash out for 45°

$v = 3{\cdot}43 \times 10^4$

$h = 15$ cm

$\sin \Phi = \frac{1}{\sqrt{2}}$

These notes in Alan Blumlein's handwriting, dated 25 September 1931,were made during the first months of his work on 'binaural', or as we know it today stereo. The notes were written in pencil in a standard notebook which is still kept at the EMI central archives. The work outlines for the first time the calculations of the principle element 'k' which Blumlein explains in his later patent No 394,325, is the phenomenon by which human hearing detects and determines the positions of sound sources.

Taken on 21 April 1938, this photograph is one of the very few known pictures of Alan Blumlein. It shows him addressing an engrossed audience of IEE members at Savoy Place, while he documents his, and his colleagues work at EMI on the recently inaugurated British high definition television system broadcasting from Alexandra Palace.

a cathode ray tube, a visual indicating system would be available.

By the outbreak of war in September 1939, EMI had produced the first of a series of prototype sound-detection systems. These were used extensively during 1940 and 1941.

Having demonstrated initiative for aircraft detection methods, somewhat surprisingly EMI were not told of the advances that had been made in electronic detection methods.

EMI had obvious advantages – not least of which was its extraordinary scientific team. But it seemed that the ministry of defence and many of the companies already contracted to produce military electronic hardware, did not consider EMI up to the task. After all, EMI had produced gramophone players and television sets before the war – hardly precision military specification products.

Another demonstration of Blumlein's genius would be required to prove to the ministry that EMI should be a part of the radar generation.

Detecting planes at night

During the first months of the war – often referred to as the 'phoney war' – there was much concern that Britain had woefully inadequate methods of detecting aircraft at night. Even worse, when British night fighters were finding German night bombers, they were not shooting them down often enough before they lost them again.

What was desperately needed was an airborne detection system that could find aircraft at night. This would then allow the night fighters to home in on an unsuspecting German bomber and close the range between the two aircraft to a distance adequate for the bomber to be brought down.

More information?

Robert's web site is a companion to the book and can be found at,

www.doramusic.com/blumlein

Eventually Rob plans to publish all 128 patents on the site.

Despite the incredible secrecy that surrounded the development of such systems, EMI was eventually made aware of the problem. With the company's inherent expertise, it was able to developed an airborne interception system that at a stroke not only improved upon any other method, but also brought EMI into the radar business – where it so desperately wanted to be.

During the dark months of 1940 and 1941, the war went very badly for Britain. It seemed that little could be done to stop the marauding German advances.

British morale was at an all-time low and something needed to be done to show the public that Britain was fighting back. However, with her Navy penned in the Atlantic protecting the vital convoys of food and provisions from U-boat attack, and her armies depleted on all fronts, the only method of hitting back at the Germans was through bombing.

Every night bombers flew missions to Germany. British propaganda claimed that they were hitting the enemy hard; the truth though was far different.

Bomber Command was indeed flying the missions, but without a practical system for locating the target most bombs that were being dropped were falling literally miles from their intended targets.

Radar that worked at night

What was needed was a system that could locate and detect a town from the air – regardless of weather conditions. And it had to be accurate enough to allow a bomber to carry out its mission. That radar system would eventually be known as H2S. Its development would become one of the closest guarded secrets of the war.

H2S relied on the centimetric properties of the cavity magnetron for its success. The magnetron had been invented at Birmingham University in 1940. It allowed far shorter wavelength pulses to be produced which in turn gave a clearer image on a cathode-ray tube display of the terrain below the aircraft.

EMI and Alan Blumlein were part of the team that had been contracted to develop and test the circuitry of the H2S system throughout the spring and early summer of 1942. It was during this series of tests on 7 June 1942 that the Halifax bomber carrying the only prototype H2S system crashed at Welsh Bicknor. Alan Blumlein and two colleagues from EMI, along with eight others were killed. Together they represented the core of the H2S scientific development team. That the H2S project was finished at all is a testimony to those who remained after the crash.

As had been predicted by Blumlein and others, H2S proved to be the instrument that Bomber Command needed. It allowed navigators to find their intended targets and bomb them with an accuracy never before achievable.

Radar's importance to Britain cannot be underestimated. At a time when the war was undoubtedly being lost, it provided the opportunity to fight back at an enemy that had seemed invulnerable just months before its introduction.

It has often been said that the atomic bomb ended the war – but that radar won it.

Yet no obituary followed his death

Following his death, Alan Blumlein's work was shrouded in secrecy. No obituary appeared and no tribute was given. For many years, various people promised a biography of this most extraordinary engineer, but none was forthcoming. As time passed, those who knew him personally grew old and died; and today but a few remain.

Imagine a world that did not have a record of Faraday, Whittle, Maxwell, Edison or Bell. Given time, Alan Dower Blumlein will receive the credit that he so richly deserves. It was for that reason that I wrote his biography.

Alan Blumlein is, without a doubt, one of the most brilliant engineers of the twentieth century, and one that the twenty-first century will finally recognise. ■

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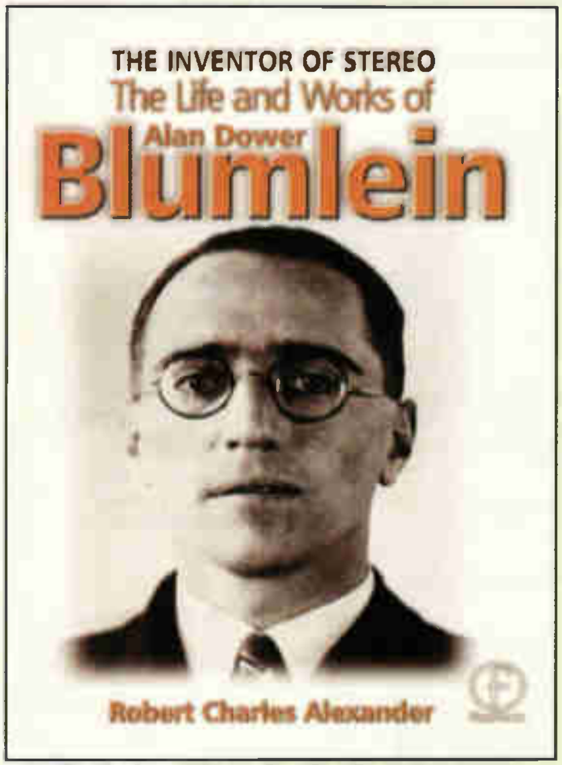

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This book is the definitive study of the life and works of one of Britain's most important inventors who, due to a cruel set of circumstances, has all but been overlooked by history.

Alan Dower Blumlein led an extraordinary life in which his inventive output rate easily surpassed that of Edison, but whose early death during the darkest days of World War Two led to a shroud of secrecy which has covered his life and achievements ever since.

His 1931 Patent for a Binaural Recording System was so revolutionary that most of his contemporaries regarded it as more than 20 years ahead of its time. Even years after his death, the full magnitude of its detail had not been fully utilized. Among his 128 patents are the principal electronic circuits critical to the development of the world's first electronic television system. During his short working life, Blumlein produced patent after patent breaking entirely new ground in electronic and audio engineering.

During the Second World War, Alan Blumlein was deeply engaged in the very secret work of radar development and contributed enormously to the system eventually to become 'H2S' - blind-bombing radar. Tragically, during an experimental H2S flight in June 1942, the Halifax bomber in which Blumlein and several colleagues were flying, crashed and all aboard were killed. He was just days short of his thirty-ninth birthday.

For many years there have been rumours about a biography of Alan Blumlein, yet none has been forthcoming. This is the world's first study of a man whose achievements should rank among those of the greatest Britain has produced. This book provides detailed knowledge of every one of his patents and the process behind them, while giving an in-depth study of the life and times of this quite extraordinary man.

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H2S - The coming of centimetric radar
The loss of Halifax V9977
Legacy
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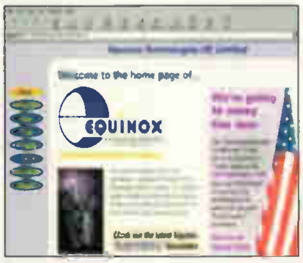

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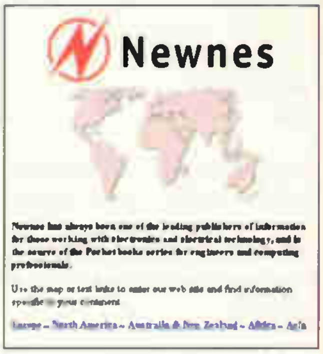
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CIRCLE NO.130 ON REPLY CARD

Continued from page 241

Analogue switch SG_1 gates the received echo to SG_2 and SG_3. The area of the echo under the early and late gates is thus integrated by I_1 and I_2 respectively. The inverted early gate output is summed with the late gate output, the result controlling the period of M_1, waveform b) of Figure 4.

Integrators I_1 and I_2 are reset by the transmit pulse tr each sweep, so that the SUM output is always updated to correspond to the current position.

Tracking is initiated by setting switch SW to A, applying a fixed voltage V_1 to the gate of the FET. Potentiometer PR is adjusted to centre the bright-up cursor, g) of Fig. 4, on the selected target, and SW returned to position B. The output of OR_2, f), gives the range in PWM form, and the negative edge detect NED output, g), in PPM form.

K. Balasubramanian
European University of Lefke,
Turkish Republic of Northern Cyprus
D96

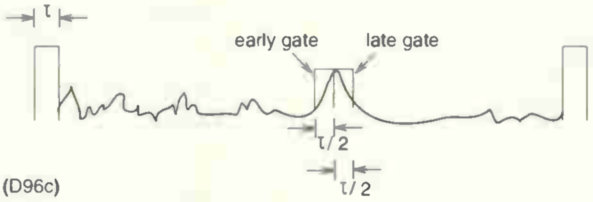


Fig. 3. Showing gates centred on target return.

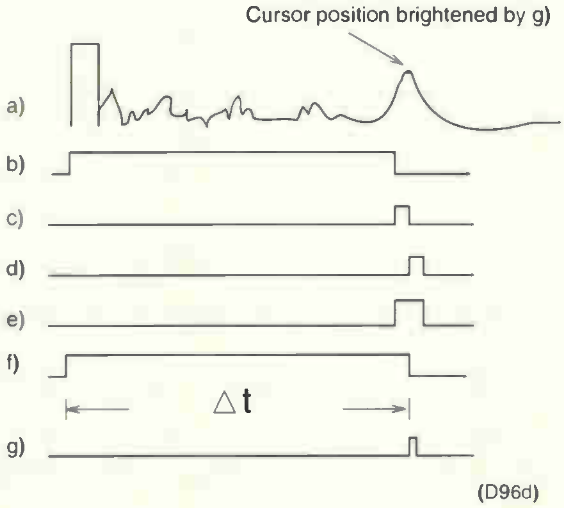


Fig. 4
a). 'A' scope trace
b) M_1 output
c) M_2 (early gate)
d) M_3 (late gate)
e) OR_1 output
f) OR_2 output (PWM range)
g) NED output (PPM range)

Latching pushbuttons for instrument control

This circuit monitors up to 16 buttons on a key panel, such as an instrument control panel, and implements the intended function of each. Any microcontroller can be used, an Atmel 89C2051 being illustrated.

The pushbuttons can be simple normally-open types, but the circuit emulates two banks, each of eight interlocked latching buttons. Each has an associated LED to indicate whether activated or not.

The buttons are arranged in two rows and only one button in any row can be active at any one time. Each pin of port 1 services a column of two buttons. The two rows are polled sequentially by the microcontroller, by driving low port 3 pin P3.4 or P3.7 to select Q_1 or Q_2. At the same time, pin P3.3 or P3.5 respectively is driven low.

A pressed button in a row is detected by the corresponding line of port 1, P1.0 - P1.7. The microcontroller writes the current state of the port in its registers, R1 for the first row, R2 for the second, and pulls down the pin with the LOW level detected, lighting the corresponding LED.

Due to the polling, any lit LED is only ON for 50% of the time, but a 560Ω series resistor usually provides adequate brightness. The 'Lock' switch disables the pushbuttons, when this condition is required.

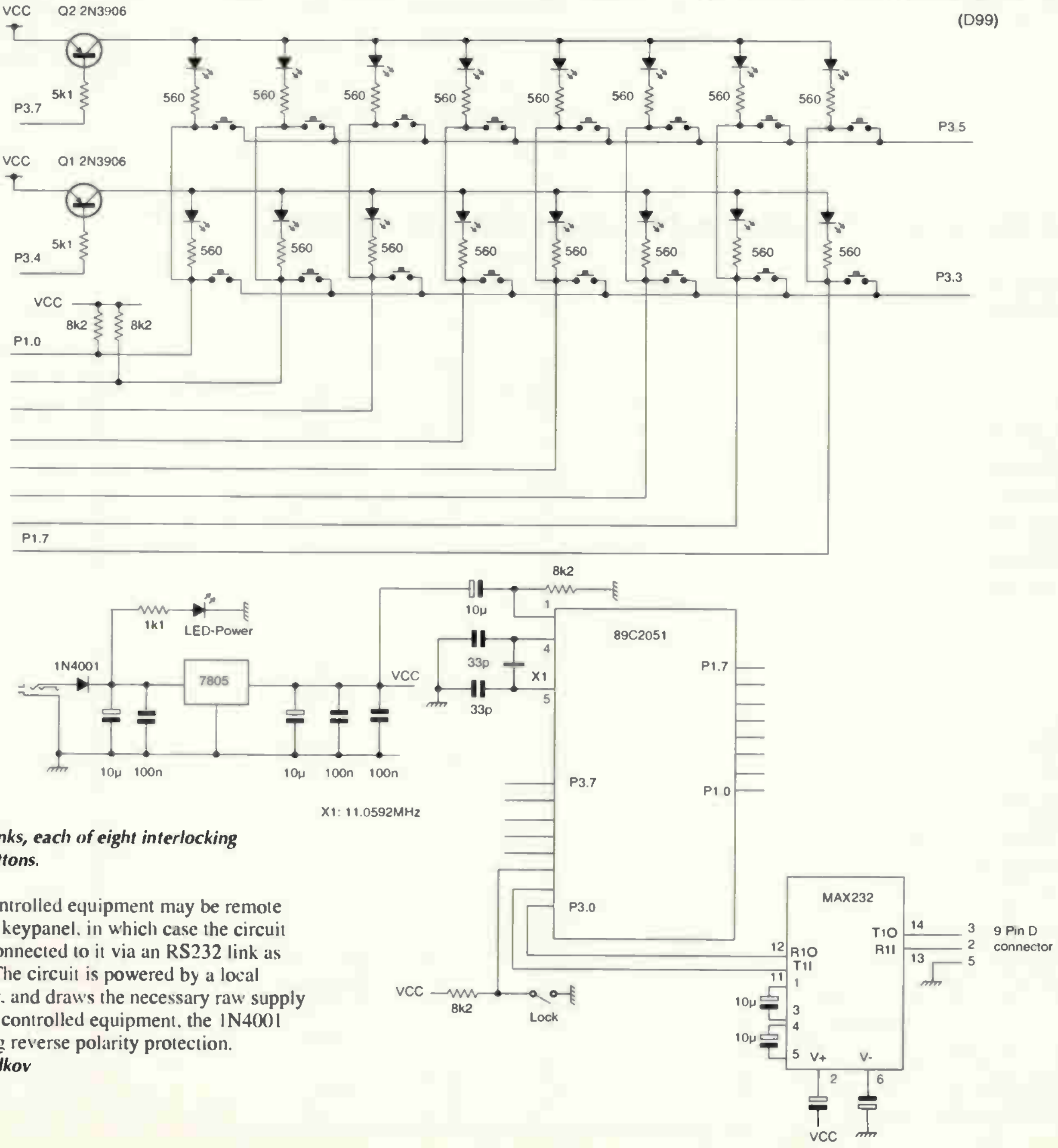
2x8 button polling program listing

```
           SETB    P3.4
           SETB    P3.7         ;Turns off the pull-ups of the two rows
           SETB    P3.3
           SETB    P3.5         ; Turns off the pull-downs of the rows
           MOV     P1,#0FFh     ;Port P1 in the input mode
           MOV     R1,#0FFh     ;R1 - memory for Row 1 - LED to be ON
           MOV     R2,#0FFh     ;R2 - memory for Row 2 - LED to be ON
Row1Check: CLR     F0           ;Checks the push buttons in Row 1
           MOV     P1,R1        ;Restores the last pushed button LED
           CLR     P3.4
           CLR     P3.3
           MOV     A,P1
           MOV     B,R1         ;Temporary LED state store
           MOV     R5,B
           LCALL   ButtonCheck
           MOV     B,R5         ;The new LED state is stored
           MOV     R1,B
           MOV     P1,R1
           LCALL   Delay        10ms delay
           SETB    P3.3
           SETB    P3.4

Row2Check: SETB    F0           ;Checks the push buttons in Row 2
           MOV     P1,R2        ;Restores the last pushed button LED
           CLR     P3.7
           CLR     P3.5
           MOV     A,P1
           MOV     B,R2         ;Temporary stores the LED state
           MOV     R5,B
           LCALL   ButtonCheck
           MOV     B,R5         ;The new LED state is stored
           MOV     R2,B
           MOV     P1,R2
           LCALL   Delay        ;10ms delay
           SETB    P3.5
           SETB    P3.7
           SJMP    Row1check
;**************************     Subroutines    *************
*****************************************************************
;******    Checks which button has been pressed - subroutine
ButtonCheck:
           JNB     P3.2,NoValidButton    ;Performs nothing if locked
           XRL     A,R5
           CJNE    A,#00h,CheckButton1
           RET                  ;No Button is pressed
CheckButton1:
           CJNE    A,#01h,CheckButton2
           CPL     A            ;Button 1 is pressed
```

```
            MOV     R5,A
;Place your button 1 action code here, F0 shows act. row
            RET
CheckButton2:
            CJNE    A,#02h,Checkbutton3
            CPL     A               ;Button 2 is pressed
            MOV     R5,A
;Place your button 2 action code here, F0 shows act. row
            RET
CheckButton3:
            CJNE    A,#04h, CheckButton4
            CPL     A               ;Button 3 is pressed
            MOV     R5,A
;Place your button 3 action code here, F0 shows act. row
            RET
CheckButton4:
            CJNE    A,#08h,CheckButton5
            CPL     A               ;Button 4 is pressed
            MOV     R5,A
;Place your button 4 action code here, F0 shows act. row
            RET
CheckButton5:
            CJNE    A,#10h,CheckButton6
            CPL     A               ;Button 5 is pressed
            MOV     R5,A
;Place your button 5 action code here, F0 shows act. row
            RET
CheckButton6:
            CJNE    A,#20h,CheckButton7
            CPL     A               ;Button 6 is pressed
            MOV     R5,A
;Place your button 6 action code here, F0 shows act. row
            RET
CheckButton7:
            CJNE    A,#40h,CheckButton8
            CPL     A               ;Button 7 is pressed
            MOV     R5,A
;Place your button 7 action code here, F0 shows act. row
            RET
CheckButton8:
            CJNE    A,#80h,NoValidButton
            CPL     A               ;Button 8 is pressed
            MOV     R5,A
;Place your button 8 action code here, F0 shows act. row
            RET
NoValidButton;                      ;No valid combination
            RET                     ;Nothing is performed
;********   Delay subroutine - generates ~10ms delay (11.0592MHz)
Delay:      PUSH Acc                ;Stores registers in stack
            PUSH B
            MOV     B,#12h
            CLR     A
Delay0:     DJNZ    B,Delay1
            POP     B
            POP     Acc             ,Restores registers from stack
            RET
Delay1:     DJNZ Acc,Delay1
            SJMP Delay0
```

(D99)



Two banks, each of eight interlocking pushbuttons.

The controlled equipment may be remote from the keypanel, in which case the circuit can be connected to it via an RS232 link as shown. The circuit is powered by a local regulator, and draws the necessary raw supply from the controlled equipment, the 1N4001 providing reverse polarity protection.

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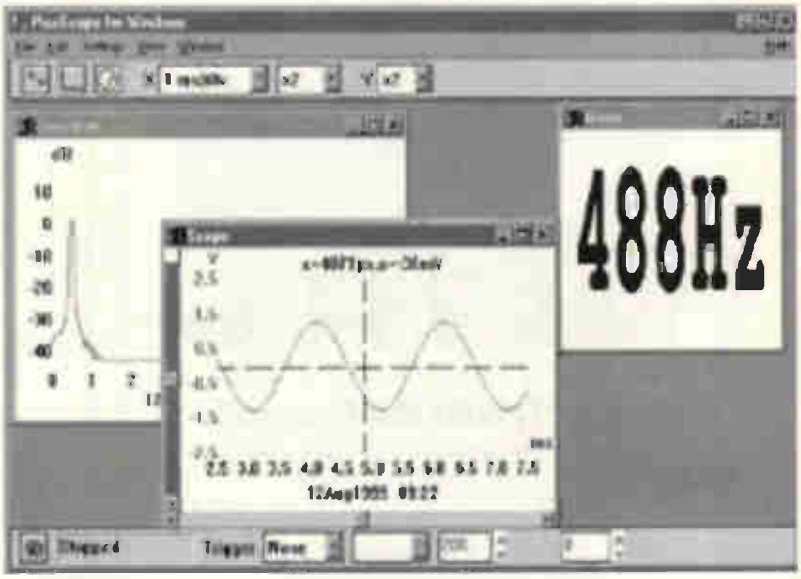

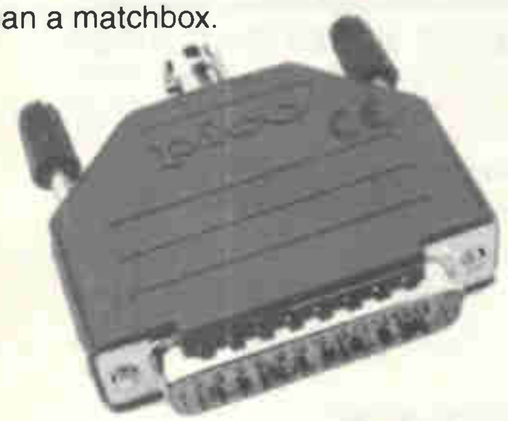

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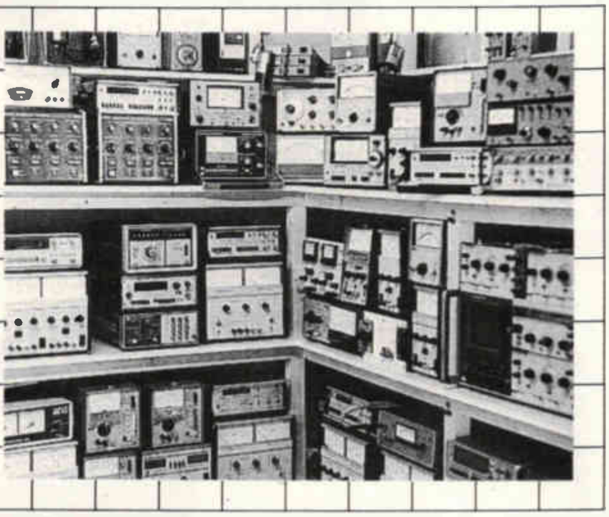

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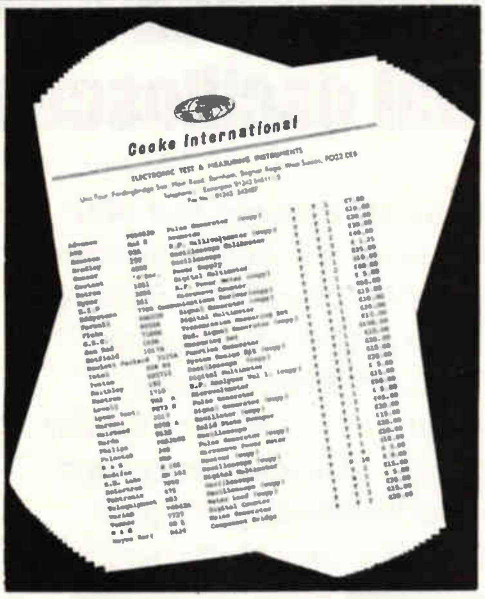

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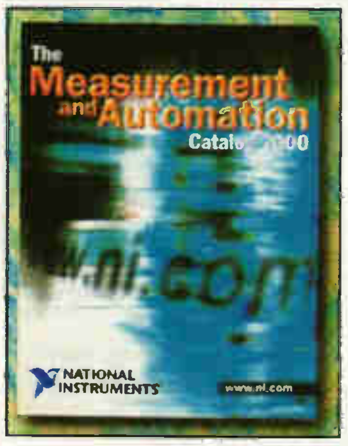

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